INTERFERENCE REDUCTION GUIDE

FOR [illegible] ENGINEERS

VOLUME 2

CHAPTER 3. CIRCUIT DESIGN

U.S. ARMY ELECTRONICS LABORATORIES

Fort Monmouth, N. J.

Prepared by:

FILTRON COMPANY, INC., NEW YORK

CONTRACT DA-36-039 SC-90707

Best
Available
Copy

1 AUGUST 1964

SUGGESTIONS OR CRITICISMS RELATIVE TO THE FORM, CONTENTS, PURPOSE OR USE OF THIS DOCUMENT SHOULD BE REFERRED TO THE DIRECTOR, U. S. ARMY ELECTRONICS LABORATORIES, FORT MONMOUTH, NEW JERSEY, ATTENTION: DIRECTOR, ELECTROMAGNETIC ENVIRONMENT DIVISION.

Qualified requesters may obtain copies of this report from DDC. This report has been released to CFSTI.

Disclaimer

The citation of trade names and names of manufacturers in this report is not to be construed as official Government indorsement or approval of the use of commercial products or services referenced herein.

THIS PUBLICATION WAS PREPARED BY
FILTRON COMPANY, INC.
UNDER CONTRACT DA-36-039 SC-90707
ISSUED BY
U. S. ARMY ELECTRONICS LABORATORIES
FORT MONMOUTH, NEW JERSEY

U.S. ARMY ELECTRONICS LABORATORIES

INTERFERENCE REDUCTION GUIDE

FOR

DESIGN ENGINEERS

VOLUME II

1 August 1964

ELECTROMAGNETIC ENVIRONMENT DIVISION
U.S. ARMY ELECTRONICS LABORATORIES
FORT MONMOUTH, NEW JERSEY

TABLE OF CONTENTS

VOLUME II

CHAPTER 3. CIRCUIT DESIGN

TABLE OF CONTENTS (Continued)

VOLUME II

TABLE OF CONTENTS (Continued)

VOLUME II

LIST OF ILLUSTRATIONS

VOLUME II

LIST OF ILLUSTRATIONS (Continued)

VOLUME II

LIST OF ILLUSTRATIONS (Continued)

VOLUME II

LIST OF ILLUSTRATIONS (Continued)

VOLUME II

LIST OF ILLUSTRATIONS (Continued)

VOLUME II

LIST OF ILLUSTRATIONS (Continued)

VOLUME II

LIST OF ILLUSTRATIONS (Continued)

VOLUME II

LIST OF ILLUSTRATIONS (Continued)

VOLUME II

LIST OF ILLUSTRATIONS (Continued)

VOLUME II

LIST OF TABLES

VOLUME II

CHAPTER 3

CIRCUIT DESIGN

Section 1. INTERFERENCE IN COMPONENTS

3-1. General

There are four basic approaches to the reduction of interference generated by such electronic circuit components as tubes (thyratrons, microwave, vhf, and uhf) and semiconductors (thermistors, transistors, diodes). These are:

1) Design of components to minimize interference generation

2) Shielding and filtering of the entire unit containing these components

3) Shielding and filtering of the components causing interference

4) Selection of circuit designs and components which create a mininum of interference

Since it is not possible to prevent the generation of some undesired signals, such as harmonics, and since the desired signals must be confined, then the components or the entire unit must be shielded to prevent radiated interference, and the leads must be filtered to prevent conducted interference. The standard, and generally most practical and economical method of shielding, is to shield the components causing the interference rather than the entire unit.

3-2. Thyratrons

a. A primary source of interference is a pulse-type plate modulator using an extremely high-voltage pulse to control the transmitter rf oscillator circuits -- especially in high-power radar equipment utilizing thyratrons for the generation of the pulse. Figure 3-1 presents a typical thyratron plate current and voltage waveform. Because these pulses constitute rapid change of voltage and current, they are extremely rich in harmonic content. A broad band of pulse-type interference can be produced by this waveform and could cover the frequency spectrum to 1000 mc even though considerably less spectrum is required. To minimize both the radiated and conducted interference, the thyratron should be provided with maximum isolation, (including efficient shielding), and circuitry that generates only the spectrum that can be utilized.

b. Thyratron tubes should be shielded and mounted on an rf isolated chassis: that is, with all leads filtered or bypassed with capacitors. Internal interference reduction circuits, such as resistance-capacitance and inductance-capacitance filters, should be installed as close to the thyratrons as possible to reduce lead length. The shield surrounding the thyratron circuit should be designed in accordance with Chapter 2, Section 4 to ensure complete rf isolation. Any lead entering the shielded compartment should be filtered at the point of entry. This includes such leads as:

1) Relay and control leads

2) Metering leads

3) Filament transformer primary leads

4) High-voltage transformer primary leads or high-voltage power leads

5) Thyratron grid input control leads

Leads are filtered most effectively by bulkhead filters that have attenuation characteristics capable of reducing the interference to low levels. The proper method of installation of these bulkhead filters is illustrated on figure 3-2.

c. The high-voltage power supply for thyratrons is also a source of interference. If the power supply is mounted within the thyratron shielded enclosure, then bulkhead filters should be installed on the high-voltage transformer primary leads; a better design, is to locate the power supply outside of the thyratron shielded area. A bulkhead-mounted filter can be designed and installed in the high-voltage line to the thyratrons for the dual purpose of smoothing the ripple frequency and attenuating the interference attempting to leave the enclosure by this path. Mounting the high-voltage transformer external to the thyratron-shielded enclosure also reduces the possibility of magnetic coupling. The transformer should be located as far from the thyratrons

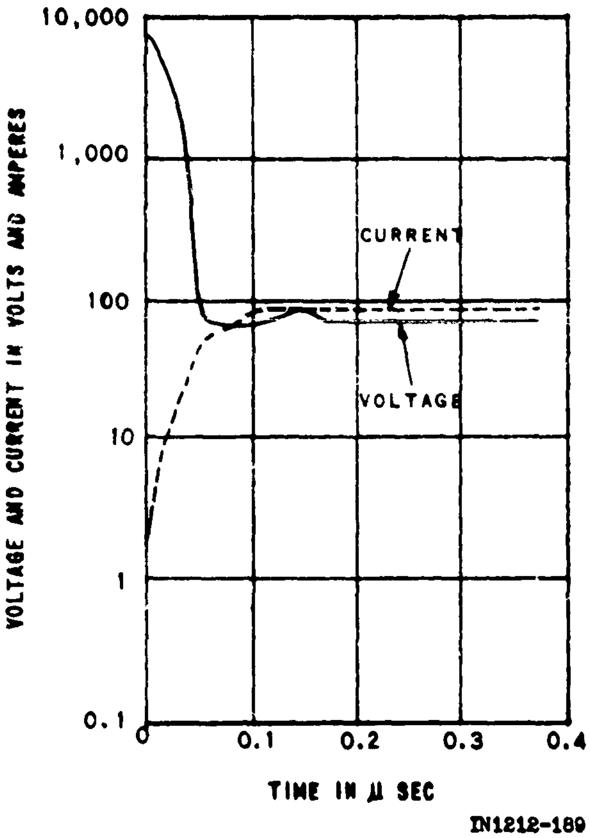


Figure 3-1. Graph of Thyratron Plate Current and Voltage Versus Time

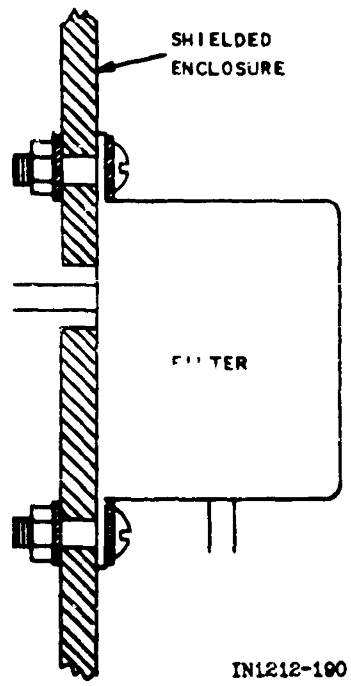


Figure 3-2. Method of Installing Bulkhead Filters

as possible, and physically oriented to minimize magnetic coupling. The trigger voltage lead to the tnyratrons should be given special consideration. This lead will conduct interference from the thyratron shielded enclosure, especially the high-voltage spike created by firing the thyratron. During establishment of a plasma in the grid-anode region when the tube starts to fire, the grid potential is raised to a very high voltage for a fraction of a μsec. This action results in a spike on the grid lead (figure 3-3). The interference action of this voltage spike may be eliminated by employing a simple, low-pass filter, such as that shown on figure 3-4. Proper design of such a filter will pass the grid voltage into the thyratron and prevent the spike from getting back into the generator. A typical design is computed as follows:

$$L = \frac{R}{\pi f_c} \text{ and } C = \frac{1}{\pi f_c R} \tag{3-1}$$

where: R = nominal terminating resistance (ohms)

f_c = cutoff frequency (mc)

L = inductance (μh)

C = capacitance (μf)

For example, if R = 50.0 ohms and f_c = 2.37 mc, then L = 6.72 μh and C = 2688 μμf. The low-pass filter should be installed in a standard bulkhead-type filter-can so that the input can be isolated from the output. A standard coaxial-type fitting can be used for the input terminal. The trigger pulse cable should be of the coaxial type. A hash filter, consisting of a choke in the plate and cathode circuits of the thyratrons, will give some reduction of interference. The use of chokes must be carefully considered because the tube life of some thyratrons is greatly shortened by the use of chokes. Even though chokes are used, interference remaining in these circuits may still be of high enough intensity to radiate from the leads; this radiation will therefore necessitate the use of shielded leads.

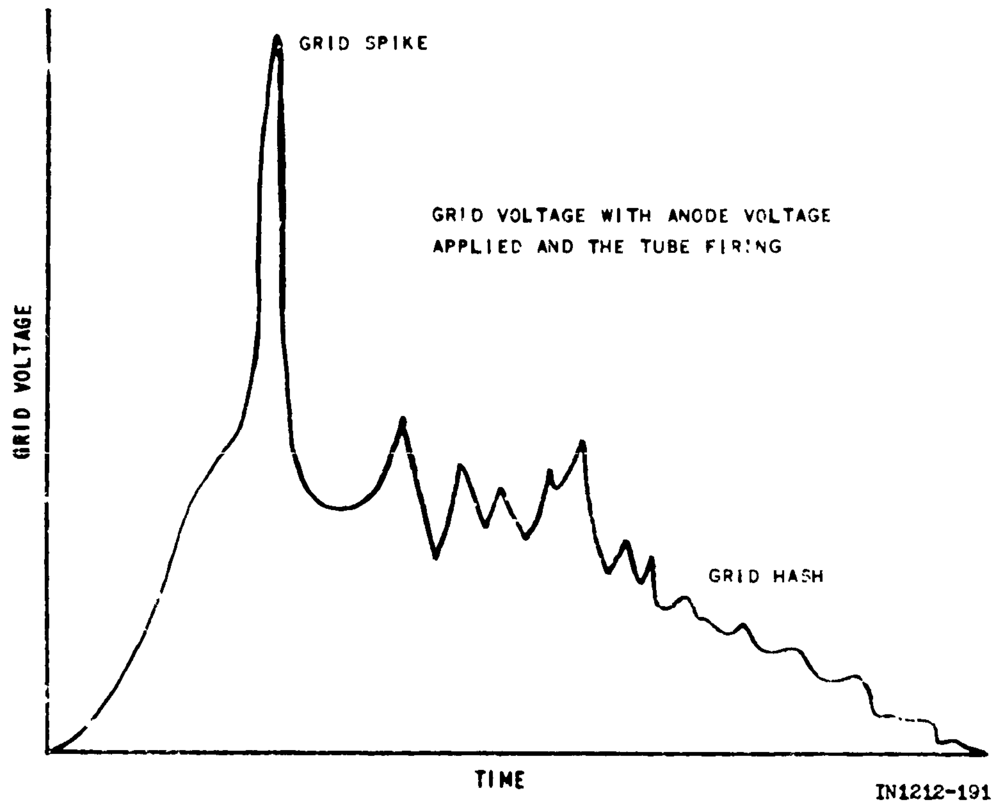


Figure 3-3. Grid Spike on Thyratron

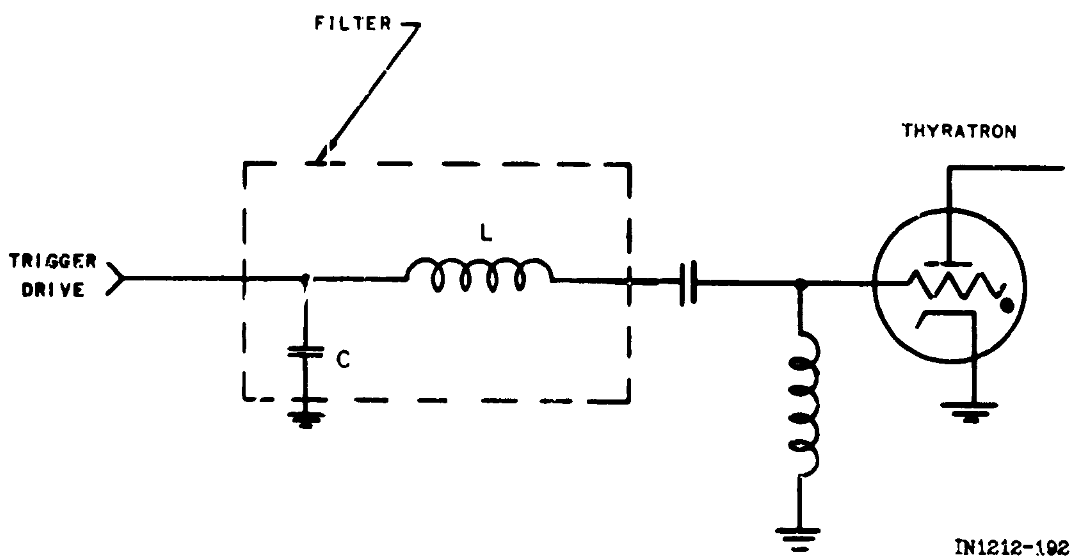


Figure 3-4. Low-Pass Filter to Eliminate Voltage Spikes

d. A shield should be constructed around the tube to prevent radiation through the glass envelope. The filament leads should be filtered at the point of entry to this shield. All leads, not having any active function in the operation of the thyratron, should be routed clear of these filament leads. If a thyratron lead does run in close proximity to leads of other equipment, it will be necessary to shield the thyratron lead that might be carrying interference.

e. A difficult interference problem arises when, for example, it is necessary to run the dc output of a thyratron to a load such as the servo motor in an antenna pedestal. A satisfactory method is to locate the thyratron(s) directly in the antenna pedestal. This method permits the dc output to be fed to the motor in a short run of conduit. The signal voltage for the thyratrons is then brought to the antenna pedestal via coaxial cables. The disadvantage of this design method, especially in a remote installation, is the inaccessability for maintenance.

3-3. VHF and UHF Vacuum Tubes

Conductors contain a large number of so-called free electrons and ions. The ions and electrons, acting similarly to an ideal gas, vibrate randomly about their average positions; these vibrations are a function of temperature. Collisions between the free electrons and the ions continuously take place, and there is a continuous transfer of energy between them. Even with no signal source applied, there are always electrons in motion, giving rise to thermal noise, a randomly fluctuating current flow. Shot noise also exists in vhf and uhf vacuum tubes. Shot noise is most commonly described as noise due to the random emission of electrons from a heated surface.

a. Inherent Tube Noise. Vacuum tubes act as noise sources because of the inherent electronic nature of their operation. Vacuum-tube noise effects include those characterized as shot effects and those resulting from such causes as tube ionization. An understanding of the mathematics of noise voltages in tubes is essential for an understanding of the possible means of reducing these noise sources.

Electrons, as discrete particles, are emitted from the cathode in a random manner; any current resulting from such emission has a random or statistical variation (shot effect). The noise for a given current is maximum when the plate is absorbing all the electrons that are liberated by the cathode; that is, when the emission is temperature-limited. If the plate does not accept all the electrons emitted by the cathode, as happens when the voltage across the tube is low enough so that not all of the electrons receive enough energy to reach the plate, there is a noise reduction. This noise reduction is due to electrons remaining in the vicinity of the cathode and forming a space-charge cloud, or virtual cathode, which serves to limit the number of electrons reaching the plate. This area is known as the space-charge-limited region. If the emitting material is irregular in its nature, then there will also be large low-frequency variations in its emissions, known as the flicker effect. Noise will also arise from variations in the secondary emission, from ionization within the tube, and from a random variation of the division of current between elements in multielectrode tubes. Of all these effects, the largest and most important source of noise is the shot effect.

b. Shot Noise.

(1) Shot effect in temperature-limited diodes. The formula for the mean-squared noise current fluctuations is:

$$\overline{i_N^2} = 2eIB \tag{3-2}$$

where e is the charge of a single electron, I is the average value of the plate current, and B is the bandwidth. This equation is valid only for low frequencies (below approximately 50 mc); the higher frequency noise phenomena involves transit-time effects that lead to increased complications. The mean-squared noise current is proportional to the plate current and the frequency bandwidth, and the rms noise current is proportional to their square root.

As an example, if I is 1.0 ma and B is 5 kc, then:

$$\overline{i_N^2} = 2\ (1.6 \times 10^{-19}) \times 10^{-3} \times (5 \times 10^3) = 1.6 \times 10^{-18}$$

The rms current generated is thus 1.26×10^{-9} amps rms. If this current flows through a 5-kilohm resistor, the rms noise voltage is 6.3 µv rms. If the bandwidth is quadrupled to 20 kc, the rms noise voltage doubles to 12.6 µv rms. Equation 3-2 has been checked experimentally many times and has even been used for precise measurements of the electron charge e.

(2) Shot effect in space-charge-limited diodes. It has been found, both experimentally and theoretically, that the random emission of electrons is smoothed out by the presence of space charge near the cathode. This is stated mathematically as:

$$\overline{i_N^2} = 2eIB\ \Gamma^2 \qquad (3\text{-}3)$$

where all the terms are the same as in the temperature - limited diode, and Γ^2 is a dimensionless space-charge reduction factor, related in complicated fashion to both the cathode temperature and the applied voltage, and varying between 0.01 and 1.0. This factor can be replaced in the equation by introducing the diode dynamic plate conductance (g_d) and the cathode temperature in degrees Kelvin (T_c) so that:

$$\overline{i_N^2} = 4kg_d\ (0.644\ T_c)\ B \qquad (3\text{-}4)$$

where k is the Boltzmann constant.

(3) Shot effect in negative-grid triodes. The negative-grid triode in the space-charge-limited region exhibits random fluctuations in its plate current. The triode mean-squared

plate fluctuation current is:

$$\overline{i_N^2} = 4k\ (0.644\ T_c) B \times \frac{g_m}{\sigma} \qquad (3\text{-}5)$$

where g_m is the triode transconductance, and σ is a constant which varies from triode to triode, normally having values between 0.5 and 1.0. As with the diode, B is the effective noise bandwidth determined by the over-all system in which the tube is connected.

c. Thermal Noise. A metallic resistor can be considered the source of spontaneous fluctuation voltages with mean-squared value:

$$\overline{V_{th}^2} = 4kTRB \qquad (3\text{-}6)$$

where T is the temperature in degrees Kelvin of the resistor, and R its resistance in ohms. The noise generated in the resistor is assumed to contain almost all frequencies and be constant up to extremely high frequencies of the order of 10^{13} cps, where quantum-mechanical effects set in.

d. Shot Noise and Thermal Noise Combined (Noise Calculations for Diode Circuits). Consider the case of a space-charge-limited diode with resistive load as shown on figure 3-5. The equivalent circuit appropriate for noise calculations is also given on figure 3-5; $\overline{i_{sh}^2}$ is the shot-noise term:

$$\overline{i_{sh}^2} = 4k\ (0.644\ T_c)\ g_d B \qquad (3\text{-}7)$$

The current version of equation 3-6 is obtained by replacing R (in this case R_L) with its conductance (G_L); $\overline{i_{th}^2}$ is the thermal noise term:

$$\overline{i_{th}^2} = 4k\ TG_L B \qquad (3\text{-}8)$$

Independent noise sources add in a mean-squared sense just as is done in signal-power calculations. It is incorrect to add rms noise voltages.

$$\overline{V^2} = \overline{V_{th}^2} + \overline{V_{sh}^2} = \frac{\overline{i_{th}^2} + \overline{i_{sh}^2}}{(g_d + G_L)^2} \tag{3-9}$$

$$= \frac{4kB}{(g_d + G_L)^2} \; (0.644 \; T_c \; g_d + TG_L)$$

It is important to stress that it is not simply a case of calculating a mean-squared thermal-noise voltage due solely to R_L, and adding to it a squared voltage term due to the diode. Each device loads down the other, and this must be considered. The load impedance seen by each current source is the parallel combination of the diode plate resistance and load resistance. Examining the last term in equation 3-9 with typical values:

g_d	$= 2 \times 10^{-3}$ mho	G_L	$= 10^{-4}$ mho
T_c	$= 1000$ °K	T	$= 293$ °K
$0.644 \; T_c \; g_d$	$= 1.3$	TG_L	$= 0.029$

The shot noise overpowers the thermal noise in this case because of the much smaller diode resistance (500 ohms) loading down the load resistance (10 kilohms). Because $g_d \gg G_L$, equation 3-9 becomes:

$$\overline{V^2} = 4k \; (0.644 \; T_c) \; r_d B \tag{3-10}$$

with B = 5 kc, $\overline{V^2} = 0.09 \times 10^{-12}$ volt^2, and the rms voltage is 0.3 μv rms.

e. Triodes.

(1) Equivalent noise-producing resistance. The equivalent noise-producing resistance of a space-charge-limited triode results from referring the triode shot noise back to the

grid circuit:

$$R_{eq} = \frac{\theta T}{\sigma T_o g_m} \qquad (3\text{-}10)$$

where: R_{eq} = that resistance which, if inserted in the grid circuit of the given tube, would produce as much noise energy as does the tube itself

θ = approximately 0.644, the ratio of the noise energy of the tube to the noise energy of a resistance (equal to dynamic tube resistance at cathode temperature)

σ = approximately 0.88, the ratio of the transconductance of the triode to the conductance of an equivalent diode (that is, the diode having a cathode-plate spacing equal to the cathode-grid spacing of the triode and having a potential of $(E_g + E_p/\mu)$ on the plate)

T = cathode temperature in °K

T_o = room temperature, °K

g_m = transconductance of the triode, mhos

In a typical triode for which $T = 1000$ °K and $T_o = 293$ °K, the use of the approximate values of θ and σ indicated, yields:

$$R_{eq} = \frac{2.5}{g_m} \text{ ohms} \qquad (3\text{-}11)$$

Equation 3-11 is the expression for equivalent noise resistance commonly used for a triode amplifier. For a triode mixer, the equation becomes:

$$R_{eq} = \frac{4}{g_c} \text{ ohms}$$

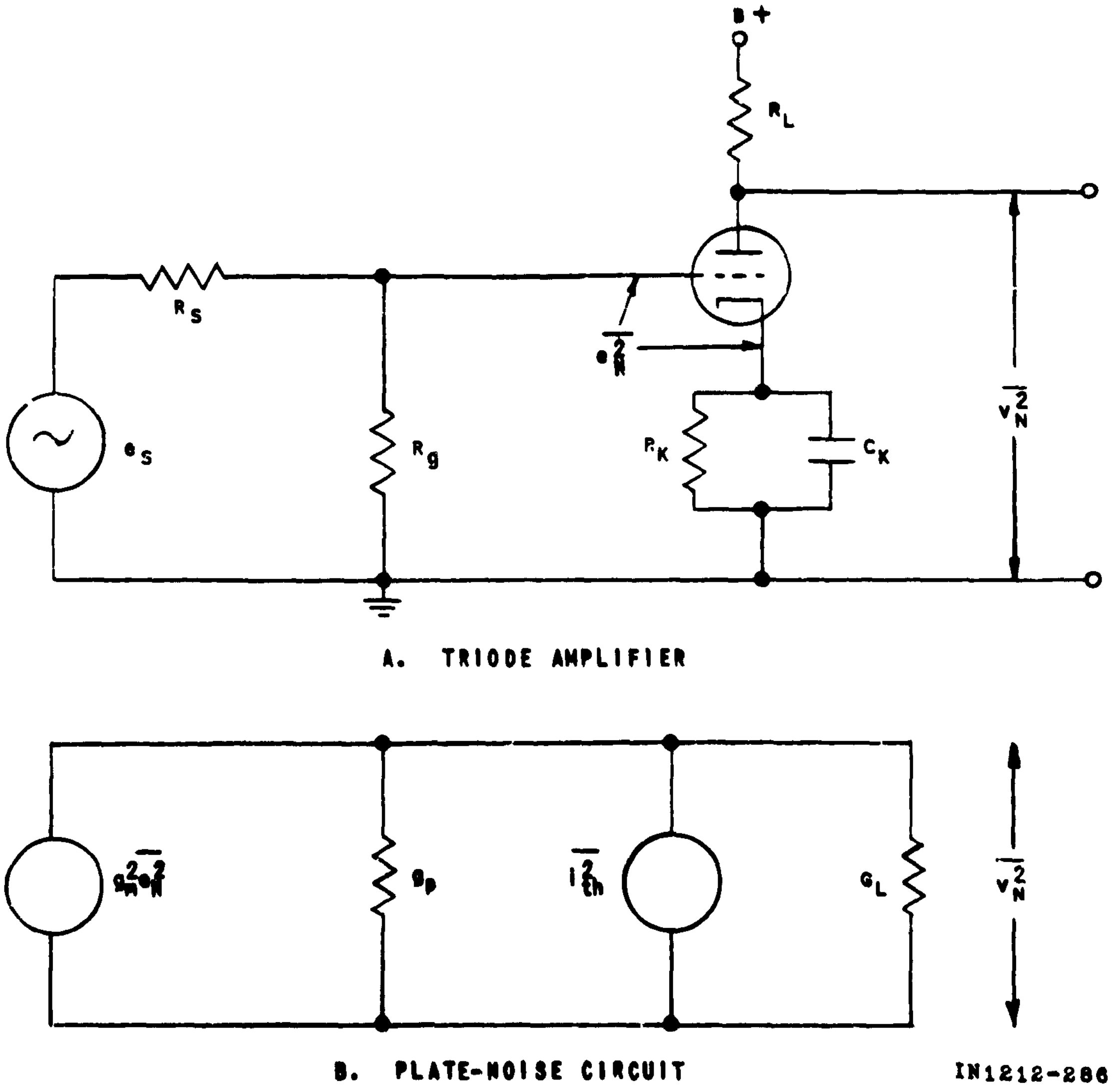


Figure 3-6. Triode Noise Calculations

where: $A_m = \dfrac{g_m}{g_p + G_L}$, the mid-band amplification of the circuit. Since the grid noise voltage appears multiplied by the circuit amplification, the thermal-noise term due to the plate load resistor is normally negligible; that is: $4kT\ G_L B \ll g_m^2\ \overline{e_N^2}$. R_L must however be included in the calculations as a resistance in parallel with r_p (the tube plate resistance).

If, for example,

R_s = 1000 ohms	$g_m = 2 \times 10^{-3}$ mho
$R_g = 10^6$ ohms	$r_p = 15 \times 10^3$ ohms
R_k = 1000 ohms	$C_k = 30\ \mu f$
μ = 30	B = 20 kc
T = 20 °C	$R_L = 75 \times 10^3$ ohms

$$R_{eq} = \frac{2.5}{g_m} = 1250 \text{ ohms}$$

The rms grid noise voltage is:

$$\begin{aligned}\overline{e_N^2} &= 4kT\ (1000 + 1250)B \\ &= (1.62 \times 10^{-20}) \times 2250 \times (20 \times 10^3) \\ &= 0.73 \times 10^{-12}\end{aligned}$$

$$\overline{e_N} = 0.85\ \mu v \text{ rms}$$

The mean-squared thermal-noise current generated by R_L is:

$$\overline{i_{th}^2} = 4kTG_L B = 0.43 \times 10^{-20}$$

But $g_m^2\ \overline{e_N^2} = (4 \times 10^{-6})\ (0.73 \times 10^{-12}) = 2.9 \times 10^{-18} \gg \overline{i_{th}^2}$

The noise from the load resistor may thus be neglected in comparison with the amplified grid noise. The rms output noise voltage is found as:

$$\overline{V_N^2} = A_m^2 \, \overline{e_N^2} \qquad (3\text{-}14)$$

$$\overline{V_N} = A_m \, \overline{e_N} = 25 \times 0.85 \ \mu v = 21 \ \mu v \text{ rms}$$

In a receiver, noise voltages are amplified just as any other voltage; therefore, the primary source of noise in a cascaded series of amplifier stages is usually the first stage. The noise introduced in succeeding stages is negligible compared with the amplified noise of the first stage, therefore, noise reduction efforts should be directed toward the first stage.

f. Multielement Tubes. The noise energy in pentode, beam, and screen-grid tubes is higher than in triodes with similar characteristics because there is an added component of noise from the random division of current between the screen and anode. The approximate value of resistance, R_{eq}, which, if inserted in the grid circuit of a pentode or similar tube, would produce as much noise energy as does the tube itself, is:

$$R_{eq} = \frac{I_b}{I_b + I_{c2}} \left(\frac{2.5}{g_m} + \frac{20 \; I_{c2}}{g_m^2} \right) \quad \text{(ohms)} \qquad (3\text{-}15)$$

where: I_b = average plate current, (amps)

I_{c2} = average screen-grid current, (amps)

g_m = transconductance of the pentode, (mhos)

The noise energy from a pentode will be about three to seven times as great as that from a triode producing an equivalent amplification. This is not at all untypical of the increased noisiness of pentode operation. There are other sources of noise in tubes, such as collision ionization, secondary emission, emission of positive ions, flicker effect from oxide-coated cathodes, microphonics and hum. These sources (with the possible exception of microphonics) produce noise that is low compared to that from the shot effect.

g. Determination of Noise Figure for Vacuum Tubes. The "noisiness" of a particular system or part thereof can be measured by comparing the signal-to-noise ratio (S/N) at output and input. This measure of the noisiness of a system is called the noise figure, F, of the system and is defined as:

$$\frac{S_o}{N_o} = \frac{1}{F}\frac{S_i}{N_i}$$

where S_o/N_o is the signal-to-noise ratio at the output, and S_i/N_i the signal-to-noise ratio at the input.

Of special interest is the amplifier problem of paragraph (e). The plate load resistor contributed negligible noise to the output; the primary source of network noise in this example, then, is the tube itself. The tube shot noise can be calculated from the equivalent grid noise resistor R_{eq}. Since the total noise appears in the grid circuit, the signal voltage for both input and output S/N is the same, and F is simply the ratio of total noise power to input noise power, or:

$$F = 1 + \frac{R_{eq}}{R_s} \qquad (3\text{-}16)$$

Electron flow acts against controlling grid voltage. The damping effect introduces a noise voltage on the grid because of the randomness of the electron flow. The magnitude of this effect varies directly with the square of the frequency, and is usually described in terms of a quantity called the transit time conductance, G_t, which must be multiplied by a temperature factor, W, when used in the noise equations. The value of W has been given as 5 for oxide cathodes. The equivalent noise conductance, G_n, is identified as being equal to WG_t. A second source of noise is the equivalent noise resistance referred to the grid, R_{eq}, as was discussed previously. Knowing these two noise parameters, the lowest possible noise figure for every frequency can be calculated, as can the source impedance at which this occurs. The formulas used to calculate the optimum source impedance ($R_{s\ opt}$) in ohms and minimum noise figure ($F_{1\ min}$) in db are:

$$R_{s\ opt} = \frac{f_o}{f} \sqrt{\frac{R_{eq}}{G_n}} \qquad (3\text{-}17)$$

$$F_{1\ min} = 1 + 2\frac{f}{f_o} \sqrt{R_{eq}\ G_n} \qquad (3\text{-}18)$$

The value of f_o is the frequency (mc) at which G_n has been determined; and f is the frequency (mc) at which the values of $R_{s\ opt}$ and $F_{1\ min}$ are to be determined. Figure 3-7 shows the predicted noise figure and the optimum source impedance of a number of uhf tube-types, as determined on small samples of commercially available tubes. Values of R_{eq} and G_n for a number of tube-types are shown in tables 3-1 and 3-2.

h. Techniques for Reducing Interference Sources. When treating a tube as a source of interference, several interference reduction - suppression and design techniques are available, as shown on figure 3-8. The techniques are:

1) Use tubes with metal envelopes or completely shield glass envelopes

2) Use Class A instead of Class B or Class C amplifiers whenever possible

3) Shield the underside of the tubes

4) Filter all bias and filament leads

5) Shield all signal circuitry

6) Use minimum power levels

7) Use good modulation techniques

8) Use amplifiers instead of oscillators at high levels

In addition, lowering the values of the noise-producing circuit elements, as highlighted in the preceeding noise voltage equations, will contribute to the over-all noise reduction.

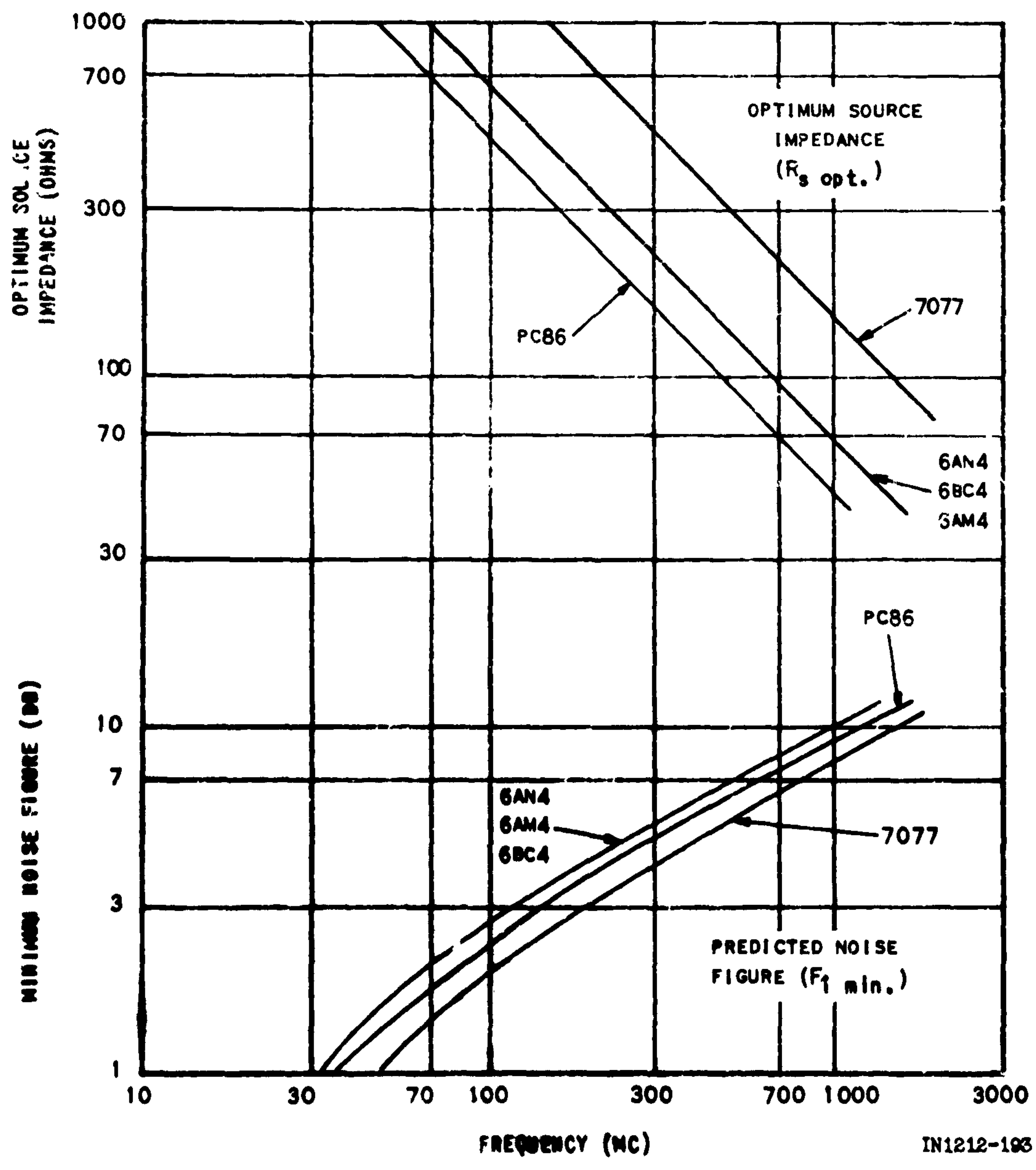


Figure 3-7. Minimum Noise Figure and Optimum Source Impedance for Several Tubes

TABLE 3-1. NOISE PARAMETERS FOR VARIOUS TUBES AT 90 MC

Tube Type	R_{eq} (ohms)	G_n (μmhos)	E_b (volts)	R_k (ohms)	G_m (μmhos)	I_b (ma)
6AM4	260	600	200	100	9,800	10.0
6AN4	250	550	200	100	10,000	13.0
6BC4	260	540	150	100	10,000	14.5
6BC8	600	320	150	220	6,200	10.0
6BK7A	240	520	150	54	9,500	18.0
6BN4	420	390	150	220	6,930	9.0
6BQ7A	435	290	150	200	7,040	9.0
6BS8	390	330	150	220	7,300	10.0
6BZ7	490	350	150	220	6,800	10.0
8CE5[a]	650	1200	200	180	5,700	11.0
2CY5[a]	525	840	125	150	6,640	10.0
6201	600	320	250	200	5,300	10.0
7077	350	140	150	82	10,000	6.5
PC88	170	710	175	125	9,800	10.0
PCC88	280	540	150	220	15,000	12.0
E180F[a]	120	1160	150	82	19,000	15.0

TABLE 3-2. NOISE PARAMETERS FOR VARIOUS TUBES AT 90 MC
(I_b = 15 MA, R_K = 68 OHMS)

Tube Type	R_{eq} (ohms)	G_n (μmhos)	E_b (volts)	G_m (μmhos)
6BC8	340	520	135	9,600
6BK7	355	580	130	8,700
6BQ7A	460	520	135	8,500
6BS8	345	530	220	9,800
6BZ7	420	540	120	8,000
6BZ8	385	700	165	10,000
PCC88	180	820	100	13,800
2CY5[a]	370	700	100	9,900
6BC5[a]	455	1500	150	9,500

[a]Pentode or tetrode measured in triode connection

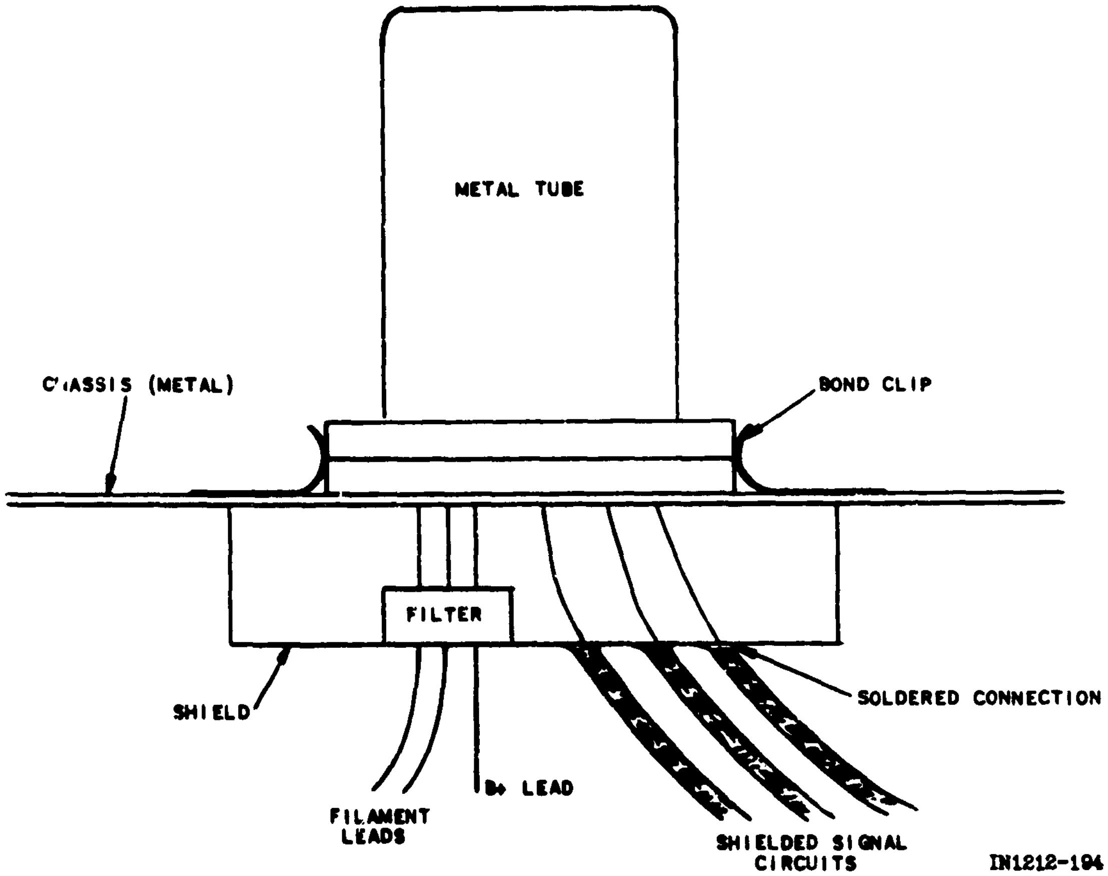


Figure 3-3. Interference Reduction Designed Tube Installation

i. Dark Noise. The use of superpower tubes in pulsed circuits raises the problem of dark noise: the emission of noise resulting from the residual plate current when the tube is cut-off. With plate voltages on the order of 25 kv, a current of only a few ma can radiate severe interference fields capable of desensitizing nearby receivers. This effect can be eliminated by pulsing the anode current, as is done in a magnetron. Tubes have recently been developed that make it possible to use grid modulation and yet hold the dark noise to a satisfactory low value.

j. Tube Susceptibility. External signals can be transferred into tube elements directly through the tube envelope or conducted into the tube by circuit leads. The degree of malfunction depends upon the tube characteristics and the relative magnitudes of the desired and interference signals. Intermodulation is often a critical problem in receivers. Its magnitude is a function of the receiver characteristics, including the choice of tubes. Proper use of tubes with high intermodulation rejection characteristics reduces intermodulation difficulties.

3-4. Microwave Tubes

a. General. High-power magnetron and klystron tubes are similar in that each utilizes a cathode, an electron beam, a microwave interaction circuit, and a beam collector. The tubes are different in the arrangement and shape of these elements and the mode of energy conversion from dc to rf.

b. Magnetrons. Present-day high-power magnetron oscillator construction usually consists of a cylindrical structure containing a circular cathode, an electron beam, and a series of coupled cavities grouped in a circle around the center cathode. The coupled cavities form a slow-wave interaction circuit and are normally strapped together for attainment of better rf performance. These cavities constitute a microwave band-pass filter. In the magnetron, this filter has a fixed number of filter sections, or cavities. The filter is continuous, and thereby made resonant when the input and output terminals are connected together. It is possible for the tube to oscillate at a number of undesired frequencies in addition to the desired frequency. These undesired oscillations fall into three separate categories: harmonic, spurious, and moding.

(1) Harmonic oscillations in magnetrons. On figure 3-9, the cavities in the magnetrons are shown in a pattern view approximately a quarter-wavelength deep at the operating frequency. Thus, for the fundamental frequency, the electric field intensity is a maximum at the tooth surface and zero at the shorted far end. At the second harmonic frequency, the cavity is a half-wavelength long; at the tooth surface, the electric field has a minimum value,

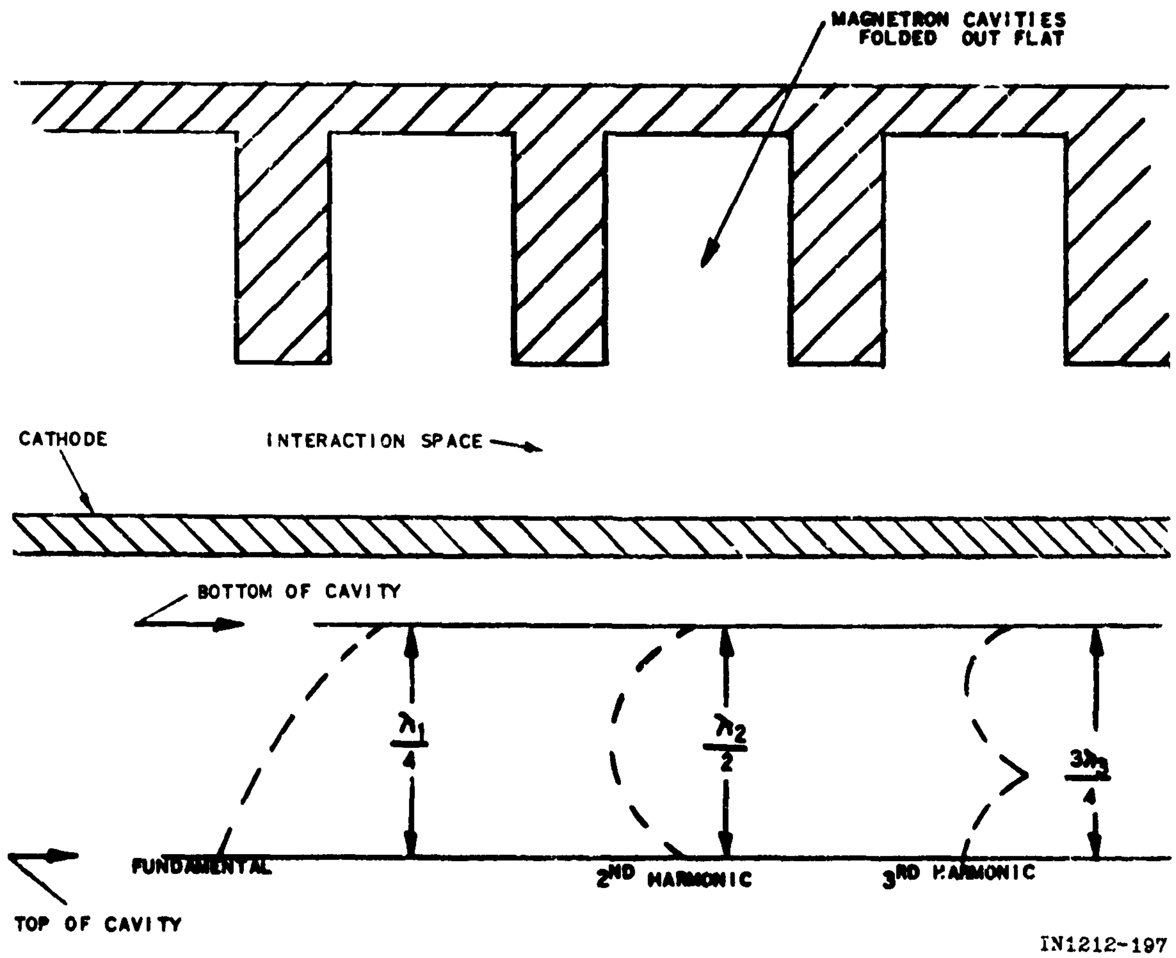
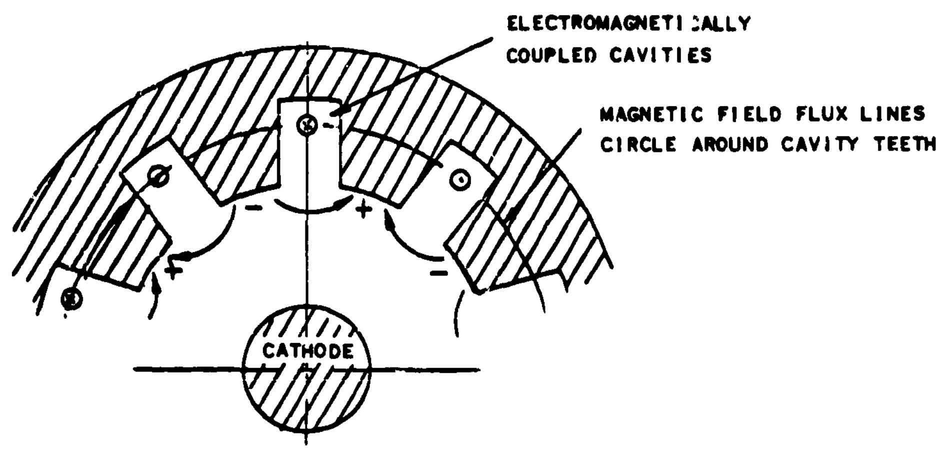


Figure 3-9. Variation of Electric Field Intensity as a Function of Distance from Top to Bottom of Cavity

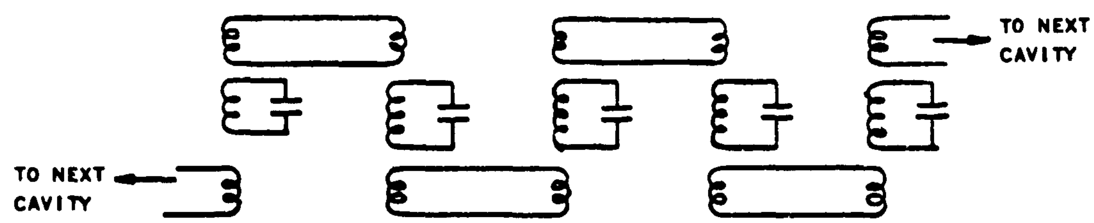
as shown. At the third harmonic frequency, the cavity resonator is three-quarters of a wavelength long, and a maximum electric field for this wave appears at the tooth surface. Even though the maximum amplitude of these harmonic fields decreases with harmonic order, the intensity of the fields at the surface of the teeth is an important factor in determining the level of harmonic energy output. For even harmonics, the rf electric field is small; for odd harmonics, it is large. Fortunately, the angular velocity of the cavity spokes is optimum only for the fundamental frequency in the tube, so that the harmonic energy output is only a small fraction of the fundamental frequency energy output. Though the level of harmonic power may be small compared with the fundamental, present-day multimegawatt units make this harmonic power a source of considerable interference.

(2) Spurious oscillations in magnetrons. If it is assumed that the resonant cavities in a magnetron are identical, then (since they are coupled together electromagnetically) the several cavities can be represented by a single tuned circuit, as shown on figure 3-10. This single tuned circuit is coupled to the output load. At frequencies off resonance, the tuned circuit appears either inductive or capacitive. Because the coupling iris and transmission-line circuit appear to the tube as a capacitive susceptance, an improperly shaped voltage pulse will shock-excite a spurious oscillation where the inductive susceptance of the tuned circuit resonates with the capacitive susceptance of the transmission line circuit. The cavities in practical magnetrons are not electrically indentical, and instead of a single spurious frequency, several spuriou frequencies are generated when various parts of the tube and output circuit resonate. The energy in these spurious frequencies can be high and can represent an important source of potential interference.

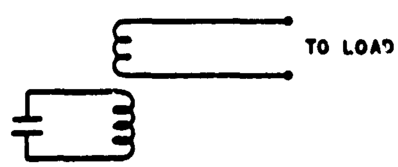
(3) Moding oscillations in magnetrons. Because a magnetron consists of n-coupled resonant circuits forming a resonant system, there are n modes of oscillation for the system. Normally in a magnetron, the design is such that the lowest frequency of oscillation occurs in the π-mode (so-called because the phase difference between adjacent anode teeth is π-radians). The frequency separation of modes in a magnetron depends upon the amount of strapping that



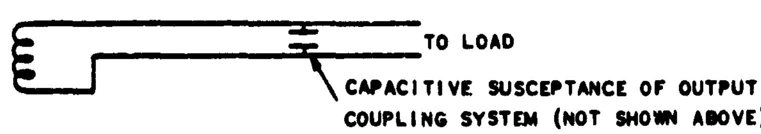
A. CROSS-SECTIONAL VIEW OF MAGNETRON OSCILLATOR SHOWING ELECTROMAGNETICALLY COUPLED CAVITIES.



B. EQUIVALENT CIRCUIT OF ELECTROMAGNETICALLY COUPLED CAVITIES.



C. SIMPLIFIED CIRCUIT FOR ELECTRICALLY IDENTICAL CAVITIES.



D. EQUIVALENT CIRCUIT AT FREQUENCIES ABOVE THE RESONANT FREQUENCY OF THE MAGNETRON CAVITIES.

IN1212-198

Figure 3-10. Evolution of Equivalent Circuit for Magnetron is applied. If jumpers with a large cross-section are used between the teeth, heavy strapping results; if jumpers with a small cross-section are used, then light strapping occurs. Figure 3-11 shows a typical magnetron with two straps, S and T. Strap S connects vanes 2, 4, and all other even-numbered vanes; strap T connects vanes 1, 3, and all other odd-numbered vanes. The cross-sectional areas of S and T are equal and determine the amount of strapping. In a strapped magnetron, the excitation of modes other than the π-mode would not normally occur if the impedance of the modulator power supply (not to be confused with the rf impedance) was well matched to that of the magnetron, so as to hold the overshoot of the pulse voltage to a low value. Often, the internal impedance of the modulator is high enough so that its load line intersects the magnetron operating points for different modes of operation. In such a case, the magnetron may mode. This condition is illustrated on figure 3-12. If the strapping rings are cut in the appropriate places, the tendency of a magnetron to mode is reduced or eliminated. The effect of cutting the strapping rings varies. In most cases, the undesired signal is eliminated; in some cases only reduced. There is no rule determining where to cut the straps; it is a matter of experience. A summary of the undesired frequencies generated by a magnetron is given on figure 3-13. The spectral density curve for a 1 μsec pulse having an amplitude of 10 to 15 kv is extremely rich in harmonics, covering the frequency spectrum from 14 kc to 1000 mc. For example, timing in the radar is dependent upon the technique of energizing the rf oscillator (usually a magnetron) with a short high-voltage pulse. The pulsing procedure leads to interference generation. The approach to the reduction of interference in the design of conventional modulator circuitry is to reduce the harmonic

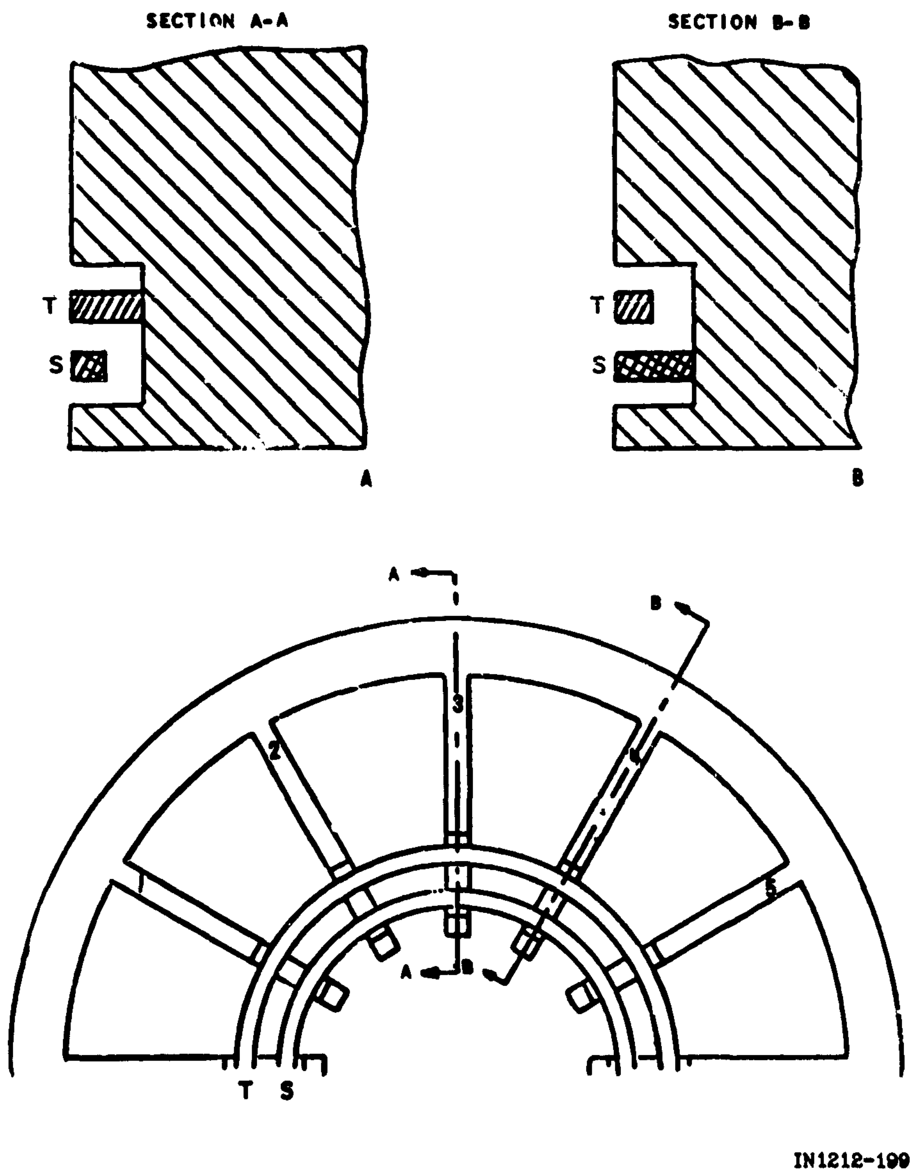


Figure 3-11. Strapping in a Magnetron Tube

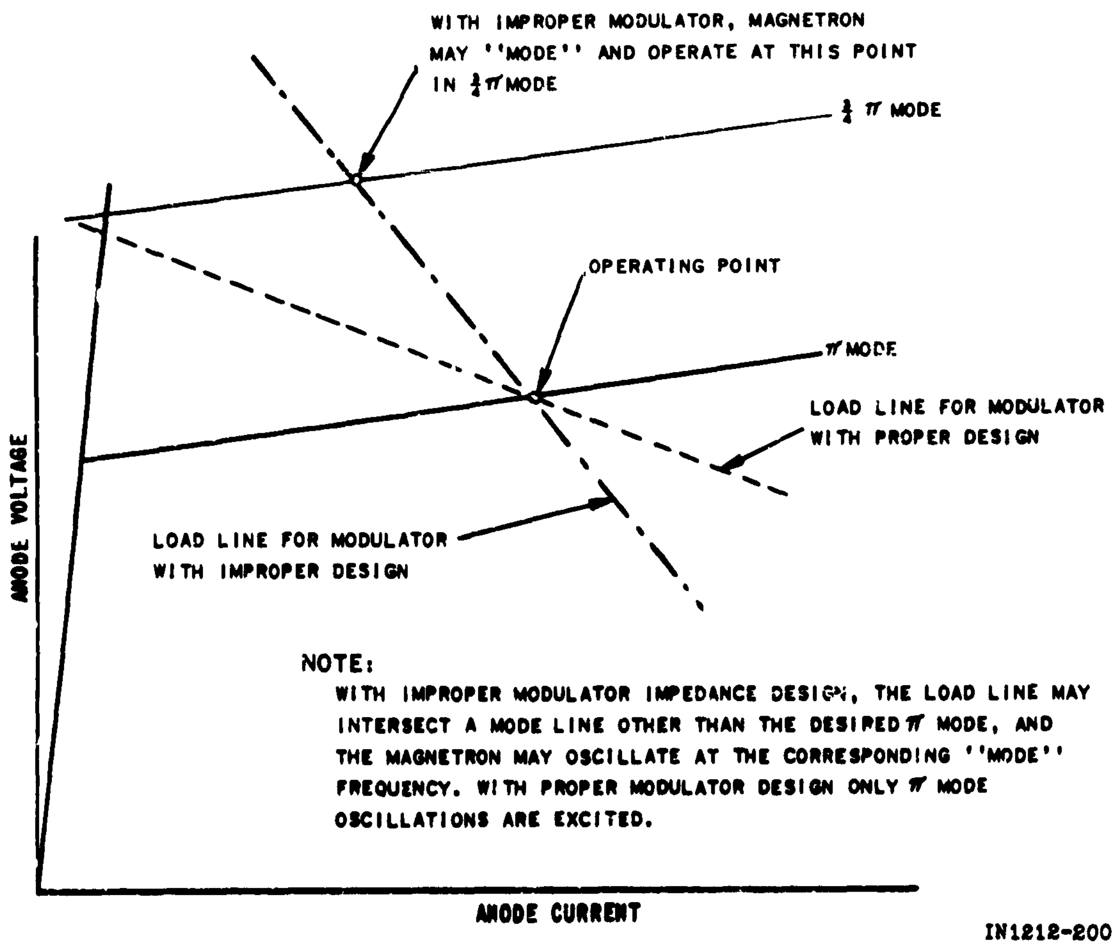
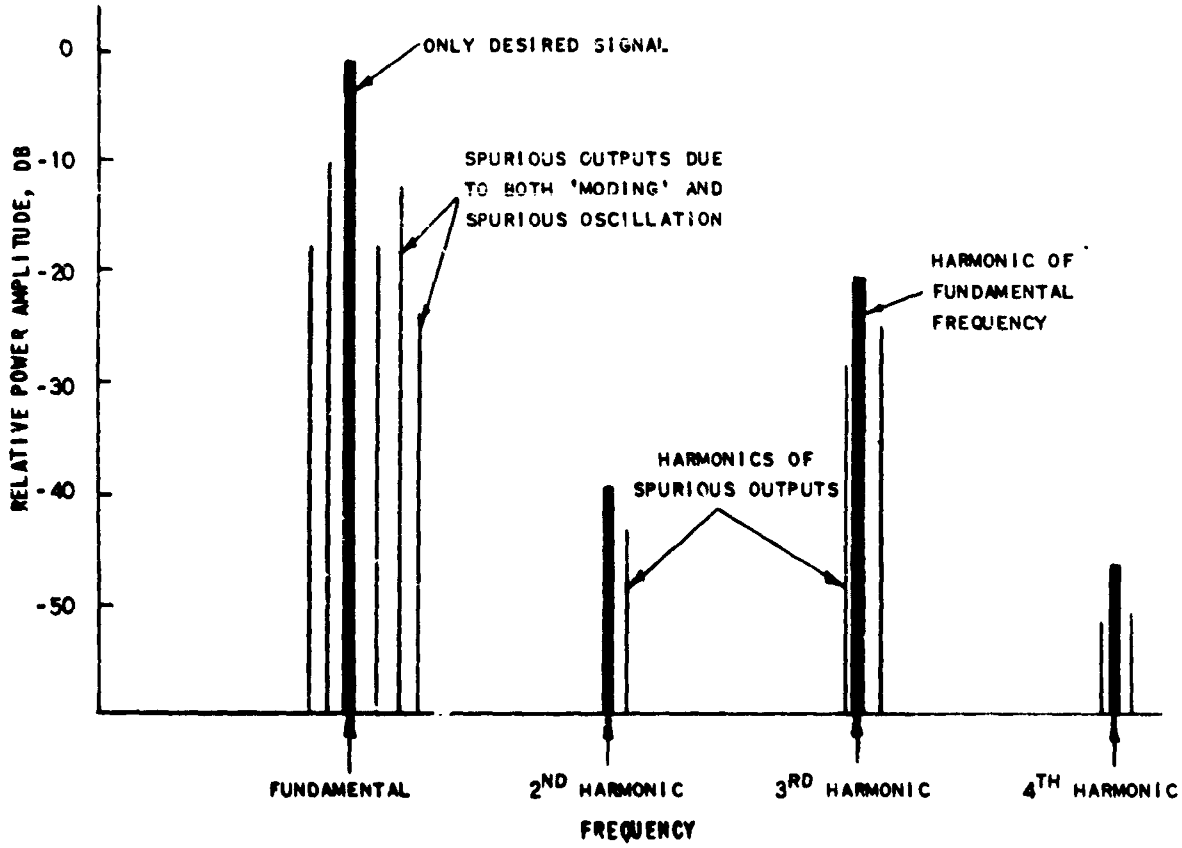


Figure 3-12. Magnetron Voltage versus Current (Not to Scale)



NOTE:
FINE STRUCTURE REPRESENTING THE $\frac{\sin X}{X}$ MODULATION PRODUCTS FOR EACH OF THE ABOVE LINES IS NOT SHOWN.

IN1212-201

Figure 3-13. Typical Frequency Spectrum of Signals Generated by a Magnetron Oscillator content of the output pulse. Sharp pulse shapes produce more interference than do more rounded pulses. Where it may be disadvantageous to destroy the pulse shape, it can be modified as a compromise between power output efficiency and reduced interference. Major interference arises as a result of the method of modulating the magnetron in radar equipment. Designs using grid modulation show appreciable merit in the reduction of interference. This technique eliminates the need for a high-voltage pulse and thus reduces the level of interference. The low-power grid-pulsing technique, used with grid-controlled magnetrons, produces some interference problems; these problems can usually be controlled by standard shielding and filtering techniques. Other advantages resulting from the use of grid-pulse magnetron modulation are reduced size, weight, and cost.

c. Klystrons.

(1) Description. A klystron amplifier may be considered as having an input cavity, intermediate cavities, and an output cavity all in a row with electron beam coupling. The input cavity serves to velocity-modulate the electron beam and to couple rf energy into the beam. The intermediate cavities and the output cavity absorb energy from the space-charge-bunching of the electron beam, and then remodulate the velocity of the beam. The rf energy of the output cavity is delivered to the output load by a coupling iris in the cavity. The remaining, weakened beam then travels on to the beam collector for dissipation.

(2) Harmonic generation. The degree of bunching of the electron beam for the fundamental frequency is a function of the distance from the input cavity to the collector. The selected position of the output cavity corresponds to the optimum bunching point for maximum

transfer of energy to the output cavity and load. It is possible for the input cavity to oscillate at a harmonic of its fundamental frequency and thus bunch the beam at a harmonic frequency. The spacing of the optimum points for the harmonic beam bunches usually is nonlinear and nonrepetitive (fig. 3-14). When the drive signal strength, or beam distance from the cavity, is changed, a change of the relative magnitude and phase of the second harmonic results. Points A and B of figure 3-14 indicate a good design location for the second cavity of the klystron. Determining a good location for the third cavity, graphically, is difficult because, at the position of the second cavity, the velocity amplitude of the fundamental (and probably of the other harmonics) changes. The change occurs because the second cavity adds its velocity modulation to that existing on the beam. It is possible to choose a location for the output cavity that would reduce coupling to the second harmonic frequency (and possibly to the third harmonic) with only a small effect on the fundamental frequency power output. Because the bunching action in the beam is inherently nonlinear, it is strongly dependent on the drive signal; hence optimum positioning of the output cavity for low harmonic output would be effective only over a narrow range of drive level.

(3) Susceptibility of klystron oscillators to interference. The signal-to-noise ratio in radar units can be affected by several parameters; one of which is the klystron reflector. The reflector is controlled either manually or by an afc amplifier, so that the frequency difference between the signal to be received and the local oscillator frequency falls within the bandpass of the if and afc amplifiers. It is conceivable that an undesired radiated signal from an external source could be picked up by the reflector's

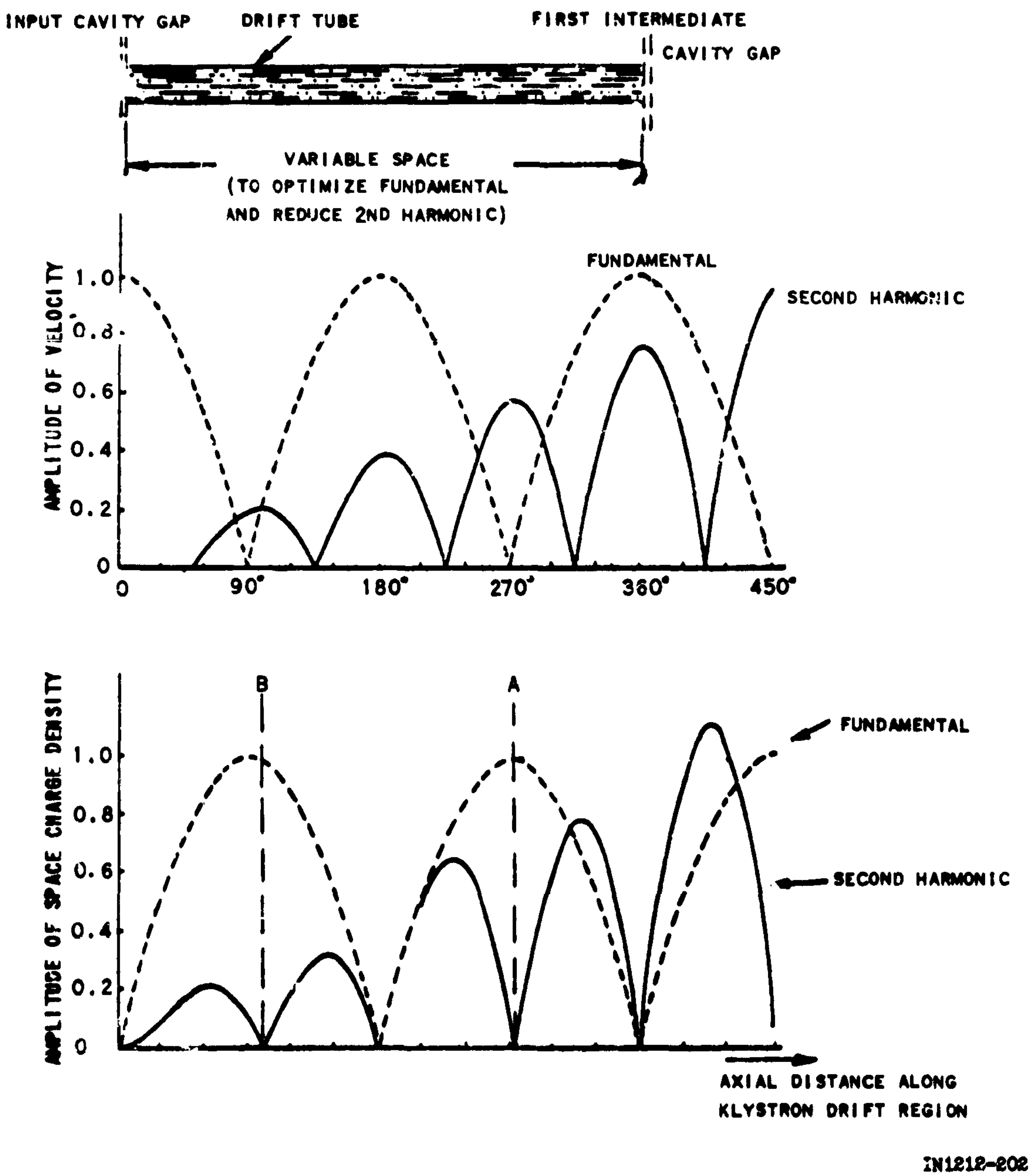


Figure 3-14. Velocities and Space Charge Density versus Axial Distance Along Klystron Drift Region for a Relatively Thin Beam

high impedance and cause the frequency of the klystron to shift to such an extent that the desired heterodyne signal would temporarily fall outside the bandpass of the if and afc amplifier. Such an occurence would result in a loss of amplitude of the desired signal and a detrimental phase shift of the rf envelope. In addition, spurious klystron frequencies, generated by interference, can mix with available ambient noise as well as with undesired signals to fall within the passband of the if amplifier.

The generation of the side bands of local oscillators can be explained in terms of modulation index, B, which is defined as the frequency change, Δf, divided by the modulating frequency, F:

$$B = \frac{\Delta f}{F} . \qquad (3\text{-}19)$$

The Bessel Function, (J), is a direct measure of relative side band strength. Table 3-3 indicates how side bands vary for values of B up to 5. For purpose of illustration, assume a cw interference signal of 2 volts amplitude, at a frequency of 2 mc, measured at the klystron reflector. If the sensitivity of the klystron reflector is 2 mc per volt, this will yield a frequency shift, Δf, of 4 mc. Therefore:

$$B = \frac{4}{2} = 2 \qquad (3\text{-}20)$$

Table 3-3 indicates that for B = 2, the first and second pair of side bands are greater than the amplitude of the carrier. The effect of these extraneous side bands mixing with system noise is comparable to an effective increase in if bandwidth and a decrease in sensitivity. As shown in the preceding material, the tendency of a klystron to produce spurious oscillations is somewhat less

TABLE 3-3. BESSEL FUNCTIONS UP TO THE SEVENTH SIDE CURRENT PAIR AND FOR A MODULATION INDCX B UP TO 5

B	J_o (B)	J_1 (B)	J_2 (B)	J_3 (B)	J_4 (B)	J_5 (B)	J_6 (B)	J_7 (B)
0.0	1.0000	0	0	0	0	0	0	0
0.1	0.9975	0.0499	0.00124					
0.2	0.9900	0.0905	0.00498					
0.3	0.9776	0.1483	0.01117					
0.4	0.9604	0.1960	0.0197					
0.5	0.9385	0.2423	0.0306					
0.6	0.9120	0.2867	0.0437					
0.7	0.8812	0.3290	0.0589					
0.8	0.8463	0.3688	0.0758					
0.9	0.8075	0.4059	0.0946					
1.0	0.7652	0.4401	0.1149	0.0196	0.0025			
1.2	0.6711	0.4983	0.1593	0.0329	0.0050			
1.4	0.5669	0.5419	0.2073	0.0505	0.0091			
1.6	0.4554	0.5699	0.2570	0.0725	0.0150			
2.0	0.2239	0.5767	0.3528	0.1289	0.0340			
3.0	-0.2601	0.3391	0.4861	0.3091	0.1320	0.0430	0.0114	
4.0	-0.3971	-0.0660	0.3641	0.4302	0.2811	0.1321	0.0491	0.0152
5.0	-0.1776	-0.3276	0.0466	0.3648	0.3912	0.2611	0.1310	0.0534

than that of a magnetron because of the nature of the coupling between the cavity and output circuit. Figure 3-15 illustrates the output of a klystron transmitter. Harmonic outputs are present at substantial power levels, while nonharmonic, spurious outputs are virtually absent. The harmonic suppression afforded by the output cavity is sufficient to reduce the total harmonic power to roughly 35 db below the fundamental. The sweep voltage from the pulse-forming network to the klystron reflector should be carried by coaxial cable to reduce the possibility of radiated interference. The klystron reflector itself is very susceptible to energy at frequencies that can result in frequency modulation of the klystron and give rise to spurious radiation. It is recommended that a small value of feed-through capacitance (0.05 μf) be shunted from the high-impedance reflector input to ground. Power supply leads should also be decoupled at the klystron. A typical installation is shown on figure 3-16.

(4) Test of klystron amplifiers. Experimental measurements were made on a four-cavity klystron, type VA 87-B. Figure 3-17 shows a plot of the total second harmonic power output (the sum of the modal powers that can propagate at the second harmonic) for this tube when operated at a fixed frequency with a varying beam voltage level. At each voltage, such items as the tuning procedures and drive level were the same; and the magnets were adjusted to focus the beam when an input signal of 2.5 watts was applied. Figure 3-17 indicates that, between the 100 kv and 90 kv levels, a change in the second harmonic level of about 3 db occurs. As the beam voltage is further reduced, another peak in second harmonic power output is observed. The normal operating beam potential for this tube was chosen to be 90 kv, the value giving the minimum second harmonic output. A second test on this tube was made by leaving the beam voltage

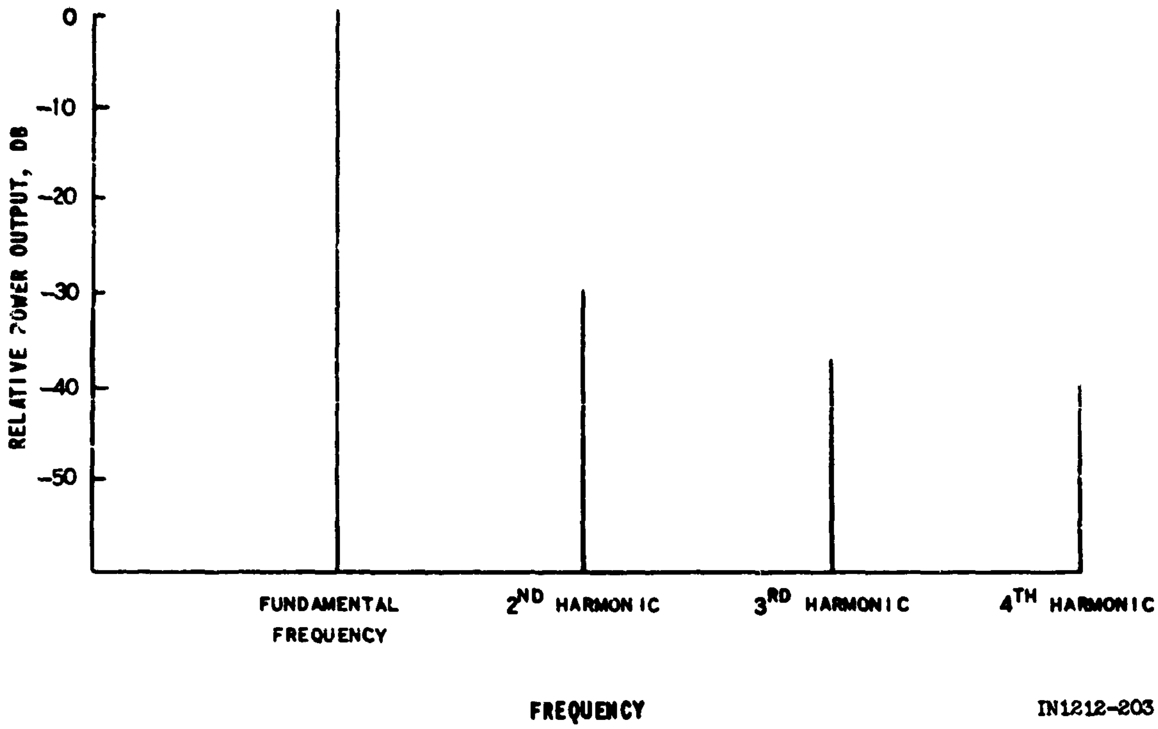


Figure 3-15. Typical Frequency Spectrum of Signals Generated by Klystron Transmitter System with Single Frequency Driving Source

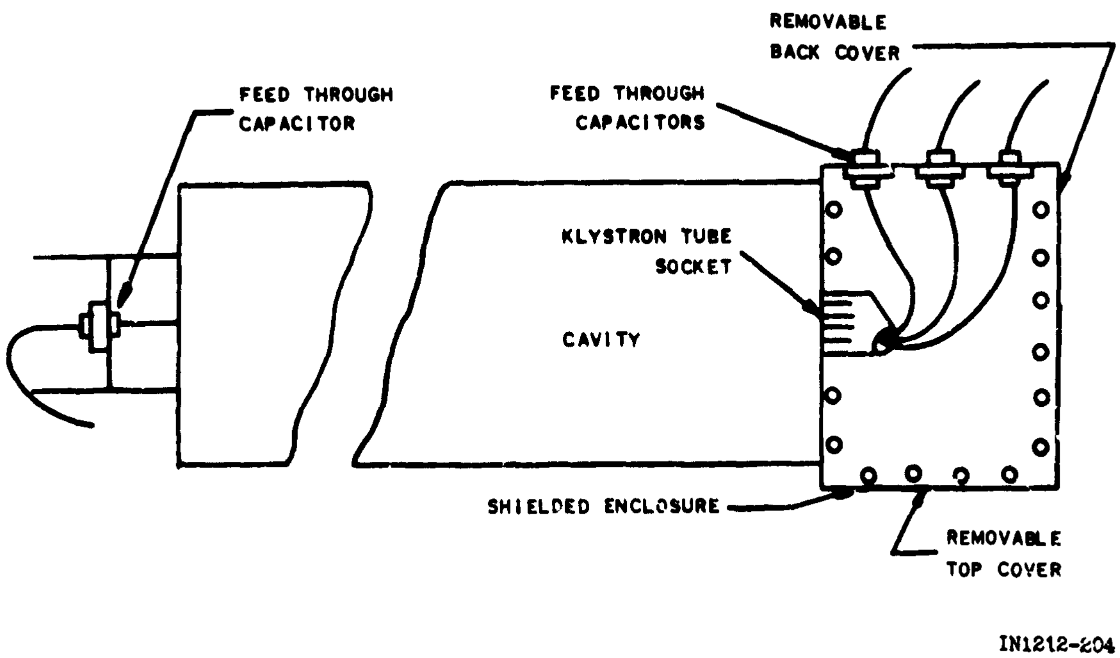


Figure 3-16. Klystron Interference Reduction

fixed and varying the drive frequency. Again, the tube was adjusted using the same procedure at each frequency. A plot of the results for this test is given on figure 3-18. Figure 3-18 shows a pattern similar to figure 3-17 in that the second harmonic power has a minimum value near the center of the tuning range. The effect of the output cavity on harmonic power may be illustrated by a series of tests (fig. 3-19).

(a) Low-power, high-gain test (beam spreading). The tube was synchronously tuned at 2750 mc and 90 kv beam voltage with cavities and beam focusing magnets adjusted to 0.1 watt drive. Then, the drive was increased to 0.25 watt giving an 0.8 megawatt output. Considerable beam spreading occured since the magnets were adjusted below saturation level. The total second harmonic power measured in this test was 3 kw peak; or a bit less than 12 db greater than the minimum value observed at this frequency with optimum tuning.

(b) Low-power, high-gain test (no beam spreading). The tube was synchronously tuned at 2750 mc and 90 kv beam voltage, except that the magnets were adjusted at 0.45 watt drive level, and the drive level was set at 0.45 watt for the test. The beam did not spread, and the total second harmonic measured 470 watts; or about 4 db greater than the optimum value.

(c) Overdriven case. The tube was tuned at 2750 mc with 90 kv voltage, but with a 70 watt drive signal overdriving the tube. The power output at the fundamental was 1.4 megawatts peak. The second harmonic power was 1400 watts.

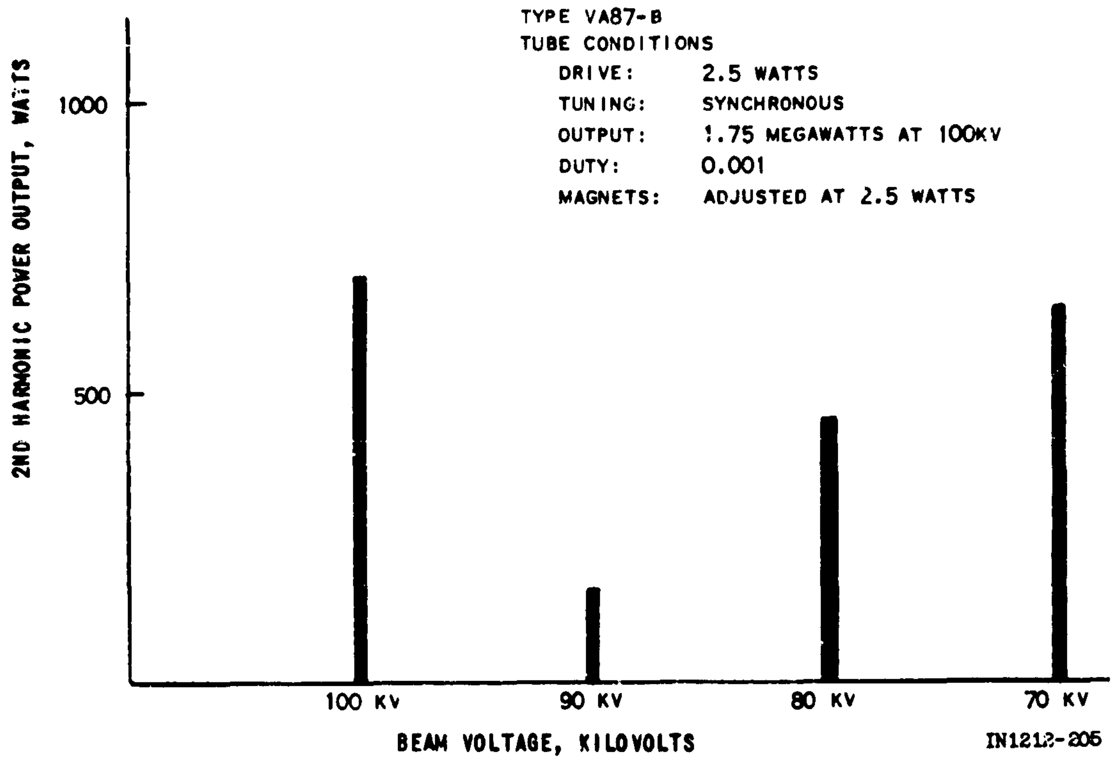


Figure 3-17. Total Peak Power Output of Klystron at Second Harmonic as a Function of Beam Voltage

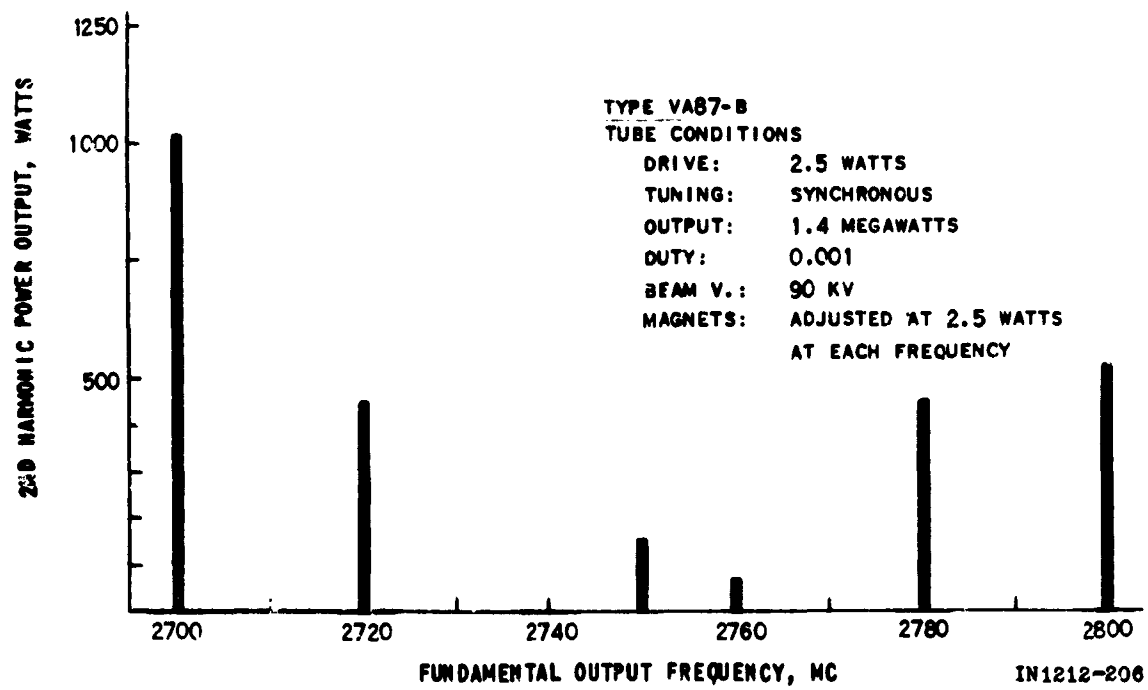


Figure 3-18. Total Peak Power Output of Klystron at Second Harmonic for Various Fundamental Output Frequencies

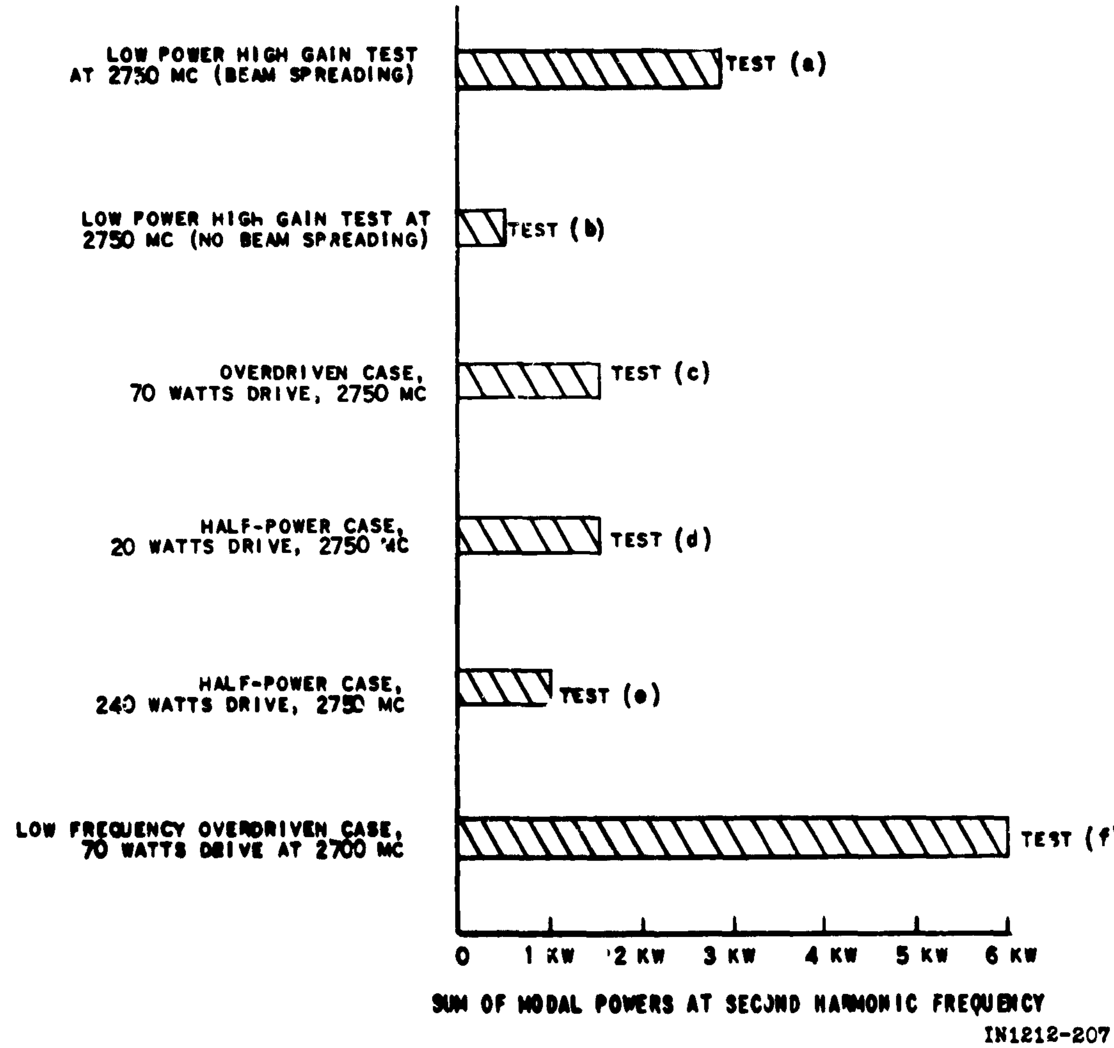


Figure 3-19. Summary Chart, Klystron Second Harmonic Investigation

(d) Half-power, case 1. The tube was tuned at 2750 mc and 90 kv as in (c), with 70 watts drive. The drive signal was then lowered until the fundamental output power was 0.7 megawatts; or 3 db lower than in (c). This occurred at 20 watts drive. The second harmonic power totaled 1600 watts; not significantly different from case (c).

(e) Half-power, case 2. The tube was tuned again as in (c), but the drive signal was increased to 240 watts to reach the second 3-db point on the overdriven side of the power output curve. In this case, the total second harmonic power level was 1.0 kw peak.

(f) Low-frequency overdriven case. The tube was tuned to 2700 mc with a beam voltage of 90 kv. At this frequency, the tube, under normal drive and tuning conditions, had a peak output at the second harmonic of 1100 watts. With normal tuning, but with the drive signal increased to 70 watts, the fundamental power output was 1.4 megawatts, with a total peak power at the second harmonic of 6 kw.

(g) Field tests. Field tests indicate that harmonic frequencies are present in rich abundance and at substantial power levels, but nonharmonic spurious frequencies are usually absent. This is illustrated on figure 3-15 in a plot of the output of a typical klystron transmitter.

d. Traveling-Wave Tube Amplifiers.

(1) Description. The traveling-wave tube, because of its peculiar power-gain characteristics, can act as limiter to protect sensitive succeeding stages. As input power to the tube increases beyond the point where power output is no longer a linear function of the input, the gain begins to decrease. At very high input power, the tube becomes an attenuator, with an attenuation value approaching that of its cold insertion loss. Generally, TWT amplifiers are wide-band, high-gain devices. Gains of approximately 30 to 40 db are achieved for small signals. Saturation power output of 0 to 50 dbm is obtained in the case of low-noise tubes. Where wide-band operation is not required, amplification of 60 to 70 db can be obtained. These high gains necessitate

that care be taken so that power lines or other connections do not introduce even weak interfering signals into the tube.

(2) Spurious frequency generation. The frequency (ω) versus delay-per-unit-length (β) diagram for a delay line in a typical traveling-wave tube is plotted on figure 3-20. Figure 3-21 shows that the traveling-wave tube delay line characteristic is similar to that of a band-pass filter, a section of which is depicted on figure 3-22A. Figure 3-23 illustrates an arrangement where a band-pass structure is considered as part of a traveling-wave tube. An electron beam (symbolically) and a series of filter sections in tandem, with a coupling between each filter section, are shown. The θ versus frequency characteristic for a section of this filter is given on figure 3-22B. A signal at the cutoff frequency, f_c, produces phase shift between each filter section (or between each of the electrodes that the beam passes) of $n\pi$ radians, where $n = 1, 3, 5.....$ The number n refers to the order of the space harmonic. The operating lines of the traveling-wave tube are shown on figure 3-24. These lines correspond to different values of beam voltage applied to the tube. The normal range of operation of the tube from this plot would be over the nearly linear range from f_1 to f_2; the range from f_2 to the cutoff frequency, f_c, is unuseable because the beam velocity would have to be changed at each frequency to achieve synchronism and amplification. The tube will not operate satisfactorily at frequencies below f_1 for the same reason. The impedance that the beam sees is the impedance between points a and b of figure 3-23. This impedance (Z) versus frequency (ω) is plotted on figure 3-25, with poles at both cutoff frequencies. In pulsed applications (including high-power radar transmitters and scatter communication transmitters), spurious frequencies may be generated. For example, with a modulator pulse voltage of relatively slow rise time, as

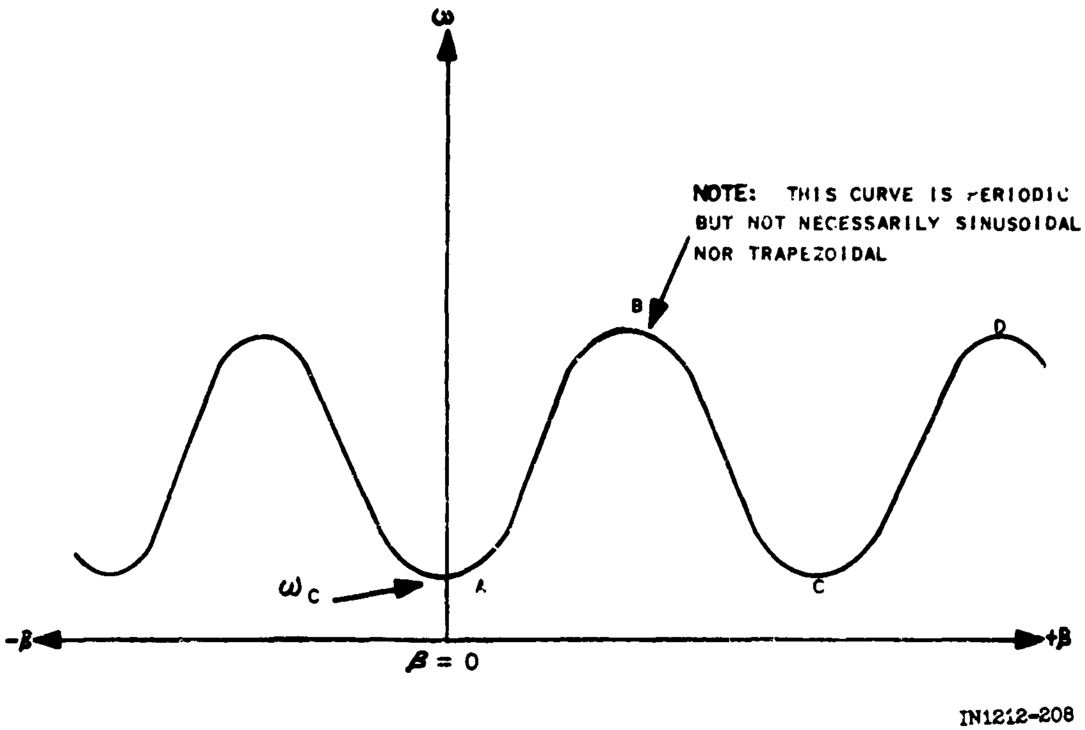


Figure 3-20. Frequency (ω) versus Delay Per Unit Length (β) for a Traveling-Wave Tube

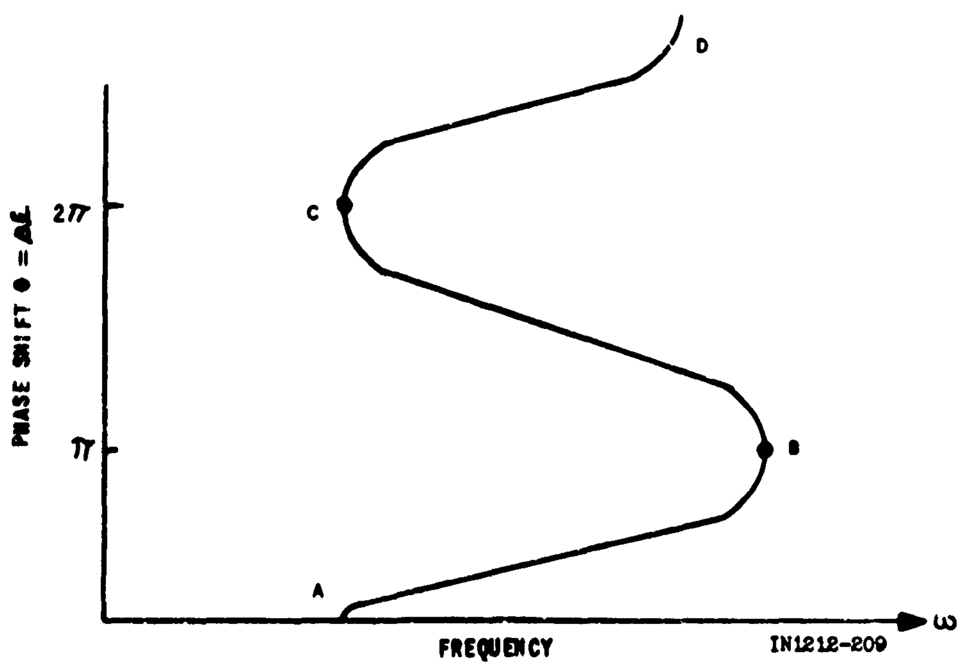


Figure 3-21. Phase Shift (θ) versus Frequency (ω) for a Traveling-Wave Tube

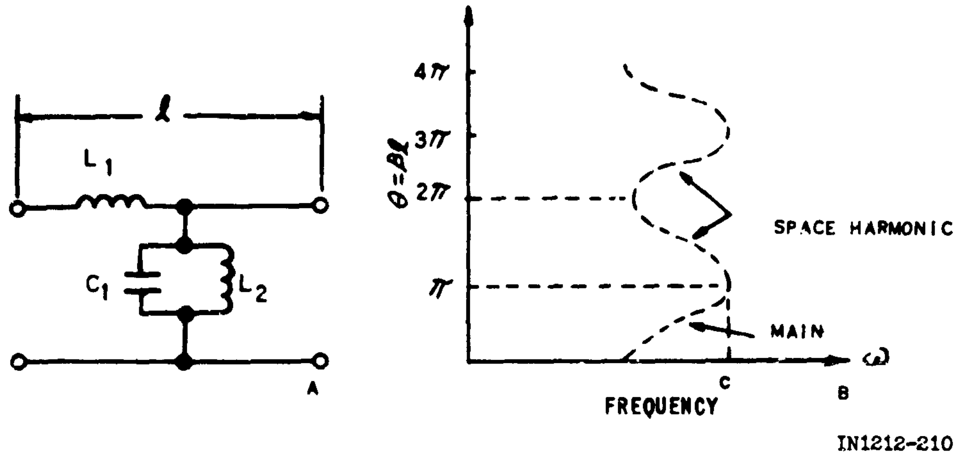


Figure 3-22. Section of Lumped Constant Band-Pass Filter with Plot of Phase Characteristic

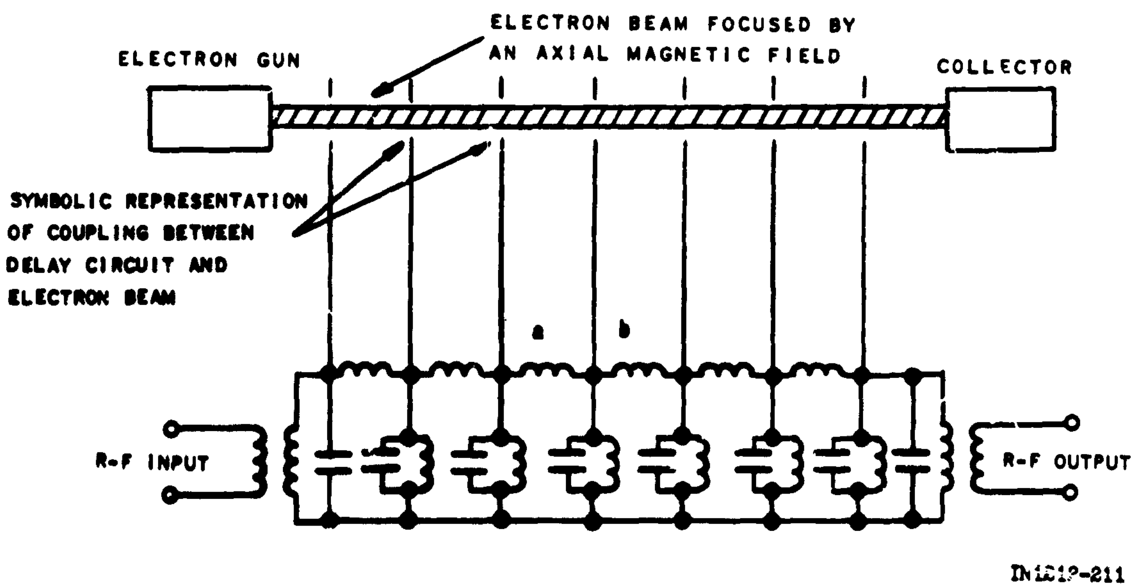


Figure 3-23. TWT with Band-Pass Delay Line Coupled Periodically to Electron Beam

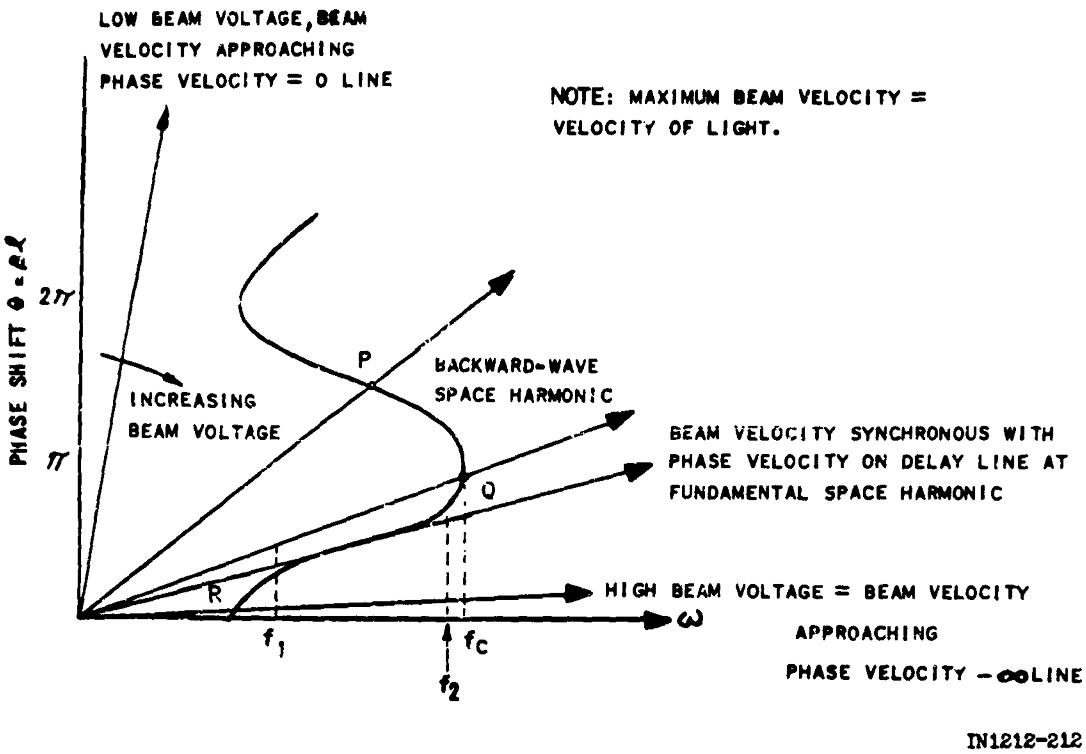


Figure 3-24. Diagram Illustrating Various Operating Voltage Conditions in a TWT

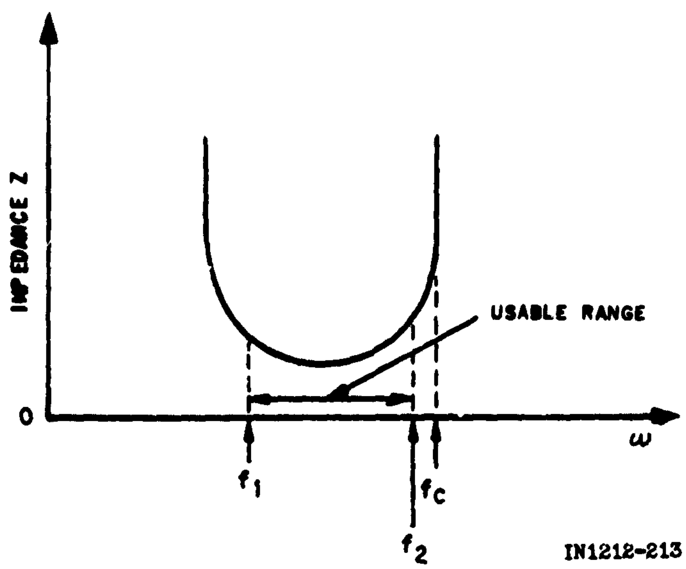


Figure 3-25. Impedance versus Frequency for TWT Delay Line

the pulse voltage rises, the line representing the velocity vector of the beam voltage rotates clockwise and passes through several points (fig. 3-24). As the line passes through point P (which corresponds to the backward-wave space harmonic) oscillations may occur; these, however, can be absorbed in the attenuator. When the line reaches point Q, the circuit impedance is very high, and slight mismatches in the rf terminal impedances on the input and output of the tube may cause oscillations to build up and give rise to a spurious frequency output. One technique used to reduce this spurious output is to increase the rise and fall times of the modulator voltage pulse. If this is done, overshoot should be avoided, or a low-frequency spurious oscillation (such as at point R) will be excited. At the low-frequency cut-off region, the delay-line circuit impedance is again high (fig. 3-25).

(3) Harmonic frequency generation in traveling-wave tubes. It is important to obtain a relationship among the relative magnitudes of the frequency harmonics in traveling-wave tubes. For a small input signal into a TWT, the power level in the second harmonic is proportional to the square of the input signal; and the power level of the third harmonic varies as the cube of the input signal. For a large input signal, for example, one near the saturation level:

$$\frac{P_{n\ max}}{P_{1\ max}} \approx \left(\frac{C_n}{C_1}\right)^3 \approx \frac{Z_n}{Z_1} \approx n\ \left[-4\pi\ r'\ (n-1)\right] \qquad (3\text{-}21)$$

where: $P_{n\ max}$ = the maximum value of the power level in the nth frequency harmonic

C_n = a gain parameter for the nth frequency harmonic

Z_n = the beam impedance for the nth frequency harmonic

r' = the radius of the helix divided by the wave-length along the helix

The equation gives the relative power of the frequency harmonic at the saturation gain point, providing the frequency harmonic lies in a pass-band in the slow-wave output coupling circuits. The relative magnitude of harmonic energy will be small if the diameter of the electron beam is held small in relationship to the diameter of the slow-wave circuit. If a hollow beam were used, the harmonic energy would be relatively large compared with a solid beam. Output characteristics of a TWT amplifier are shown on figure 3-26.

e. Backward-Wave Tubes. Backward-wave tube assemblies require an electromagneitc field for directing the electron beam. Two approaches are used for generating this field: one uses a solenoid structure axially aligned with the tube; the other uses a permanent magnet structure. Depending upon the power involved, electromagnetic fields vary in flux density over a wide range. Such fields cannot be completely contained within their operational area unless a low-reluctance path is provided for the return of the flux. This can be accomplished by mounting the structure within a magnetically-shielded enclosure. The shielding material must be of sufficient cross-section to carry the flux without any possibility of approaching the saturation level. Backward-wave tubes exhibit sensitivity to magnetic fields. The unwanted interfering fields can seriously affect the frequency stability when the tube is used as an oscillator. The shield prevents entry of external magnetic fields and contains the internally generated fields.

f. TR-ATR Tubes. Another source of interference is the TR tube, a cold-cathode gas tube associated with a resonant cavity, that functions as a switch. In radar units, it short-circuits the receiver input by ionizing during the transmit time. In most radar units, an ATR tube is used with a TR tube. Both TR and ATR tubes usually require a high keep-alive voltage to provide partial ionization between transmit pulses. The lead supplying this voltage should be shielded and

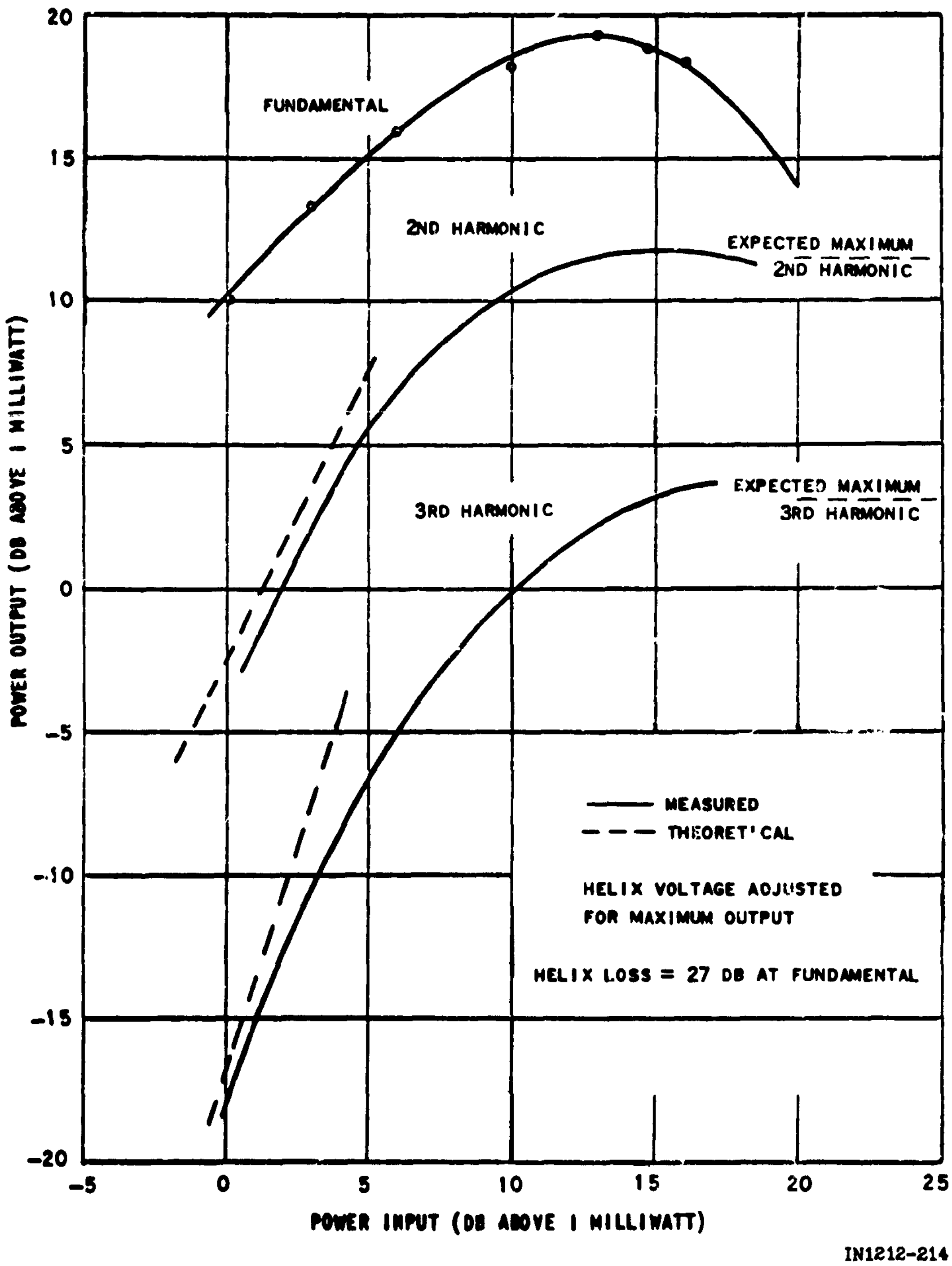


Figure 3-26. Output Characteristics of a TWT Amplifier

filtered at the tube. The capacity of the shielded lead will combine with the current-limiting resistor, when the resistor is located at the tube, to provide decoupling. When shielding is not provided for the bias lead, pulse interference, at the pulse repetition rate, may be induced in nearby wiring.

g. Tube Suppression.

(1) A useful interference reduction design technique is to use diodes to switch capacitors across tuned coaxial cavities, thus detuning the cavities and desensitizing the receiver. This action removes the need for such devices as relays or TR tubes. The technique is simple and fast acting; it is a quick and convenient way to modify a receiver that was not originally designed with adequate front-end protection.

(2) Figures 3-27 and 3-28 show a uhf receiver's front-end and its protective circuits. When a gate signal arrives, it biases diodes D_1, D_2, and D_3 in the forward direction, thus switching capacitors C_1 and C_2 across the tuning ends of the rf preselector cavities, and C_3 across the rf output cavity. This action detunes the receiver input during the desired period. The diodes are type 1N251. They have a more than adequate switching speed and operate satisfactorily in the uhf region; they also have, with an applied reverse bias, a shunt capacitance of less than 0.8 pf. During normal receiver operation, the diodes are biased in the reverse direction, and the low shunt capacitance permits tracking of the preselector and the rf stage over the uhf band. When a high-level signal arrives at the receiver input, the diodes are biased in the forward direction, desensitizing the receiver.

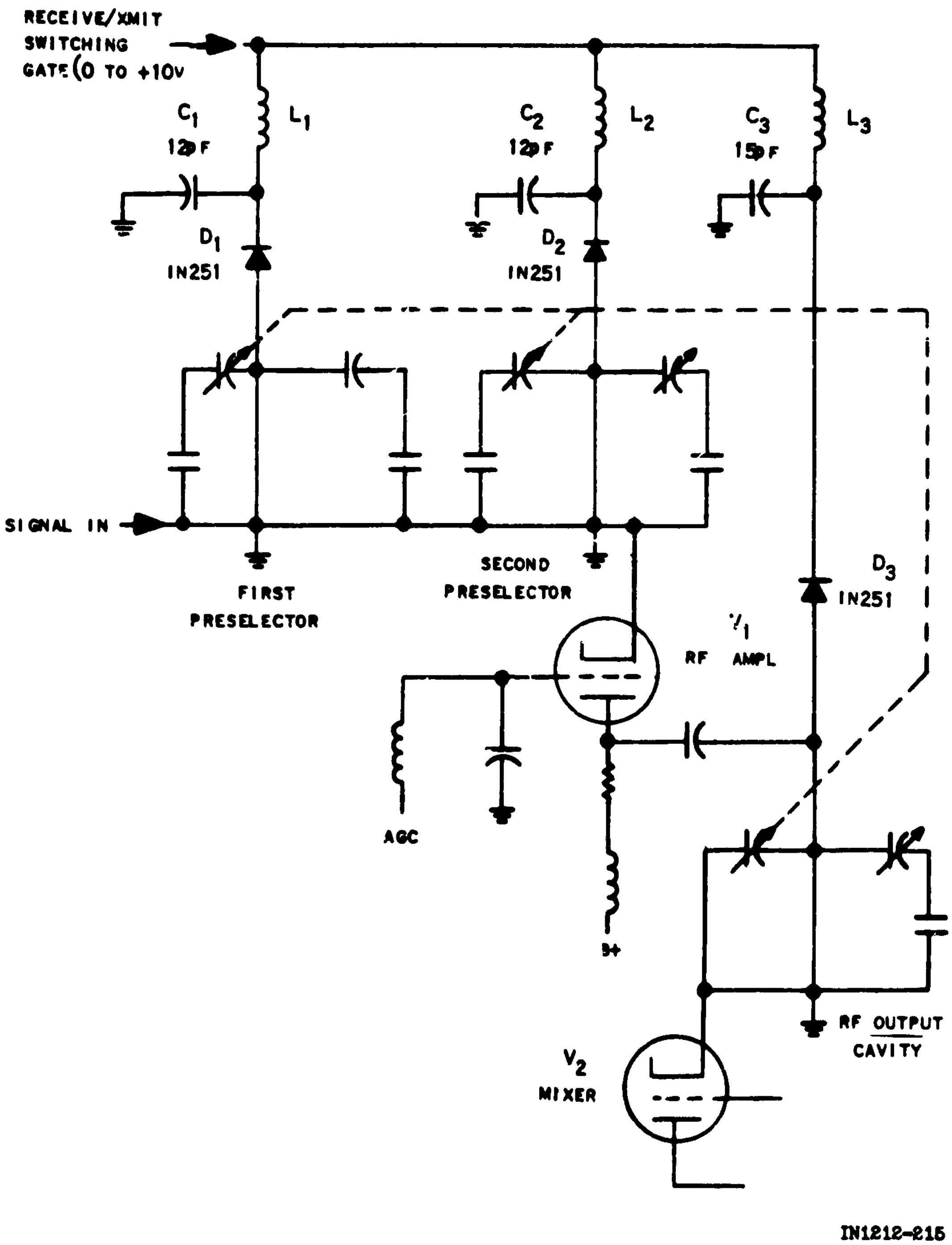


Figure 3-27. Protective Circuits and Receiver Front End

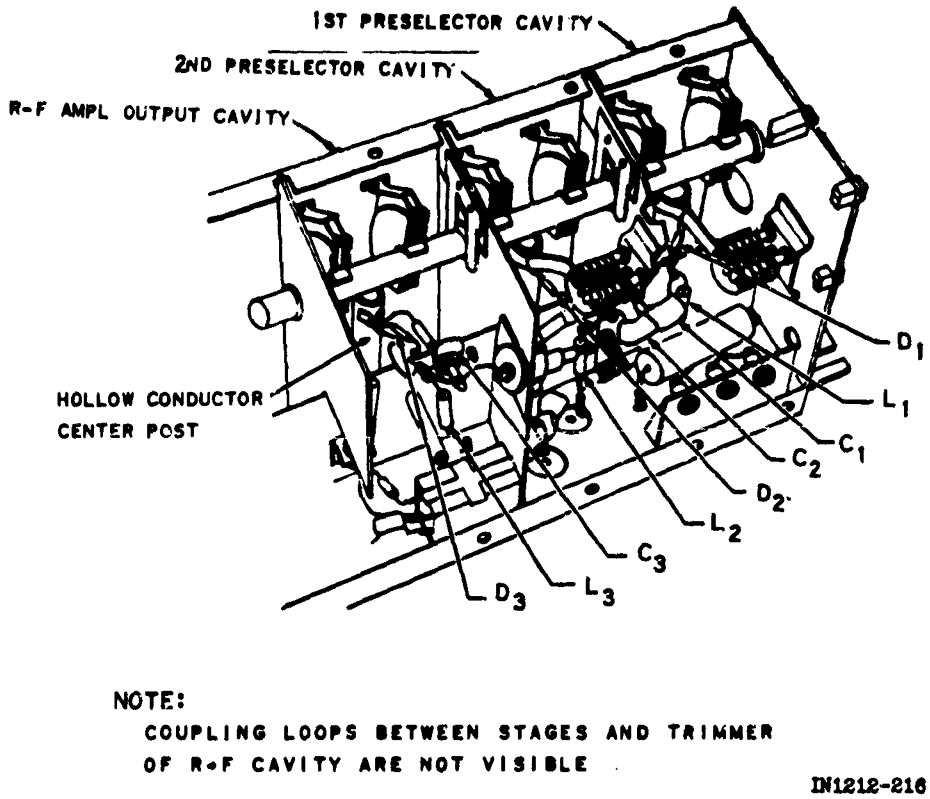


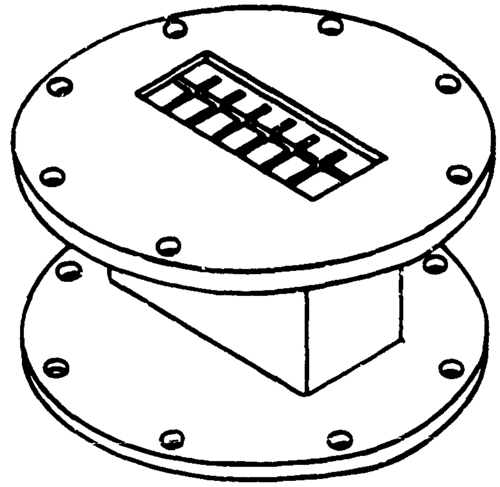
Figure 3-28. RF Stages of Receiver

(3) In the magnetron, klystron, and TWT, spurious frequency generation can be minimized by proper shaping of the leading and trailing edges of the voltage pulse. In some klystron tubes, the generation of splatter (undesired signals adjacent to the main signal) can be reduced by designing the tubes so that the control grid voltage pulse crosses the zero bias level slowly (when compared with the total rise time). This presumes a beam pulse width longer than the rf pulse width. Because driving sources used for klystrons and TWT amplifiers generate not only a desired driving signal but also undesired signals, it is important to filter the output lead to reduce the unwanted signals in the system output. The total energy output at a given harmonic frequency can be reduced by the simple addition of a coaxial harmonic filter in the output of the high-power klystron amplifier.

(4) By using an unusually strong magnetic field to focus the electron beam of a standard traveling-wave tube, the terminal noise figure has experimentally been lowered to 1.0 db. This state-of-the-art technique has provided the lowest noise figure ever achieved for a microwave tube. The 1.0 db figure was obtained by focussing the 1.07 μa beam with a field of 4500 oersteds. The tube operated at 2.6 gc. The high magnetic field dampened both shot and thermal noise.

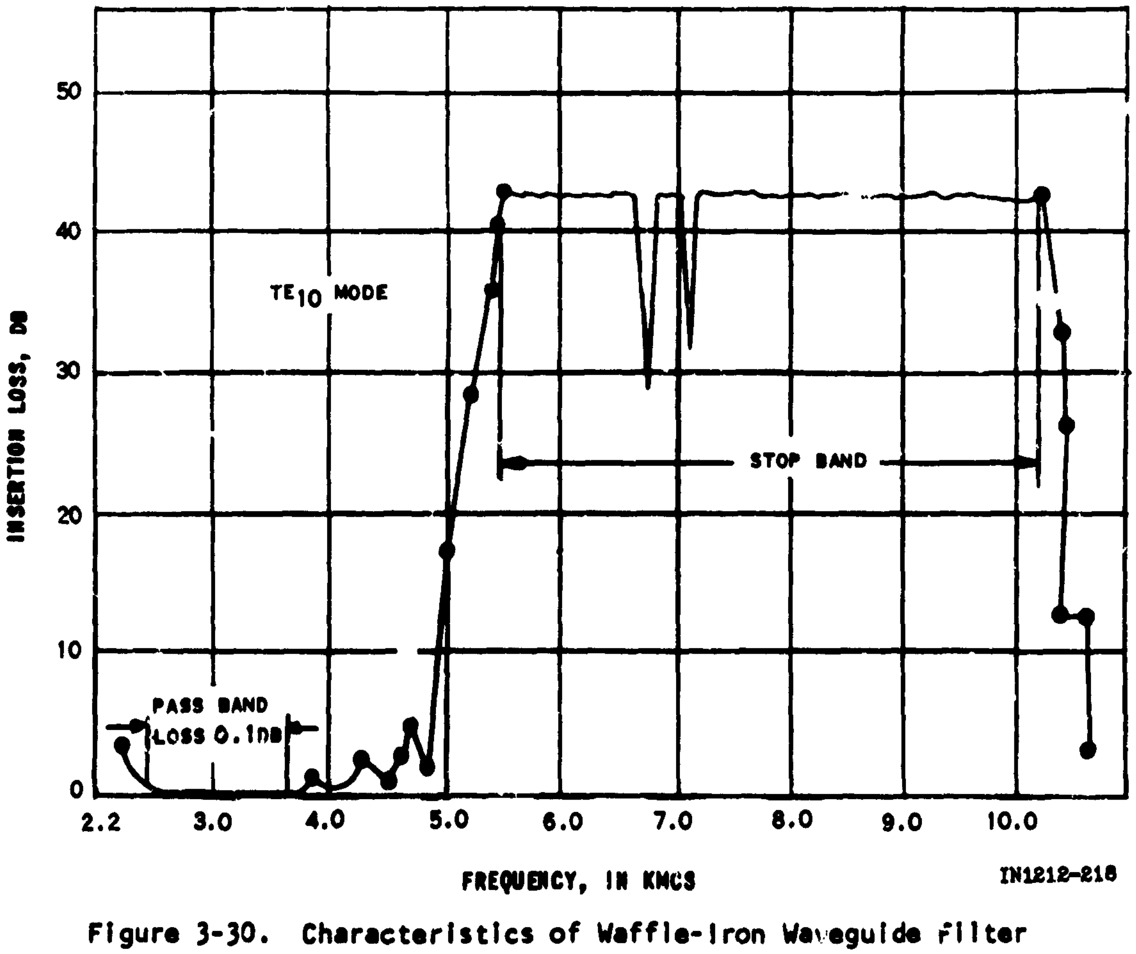
(5) One of the methods for reduction of undesired energy in microwave tubes is the employment of integral filtering within the vacuum envelope of the tube. The filtering can accomplish two objectives at the same time: it can minimize unwanted signal generation and reduce the amount of harmonic energy incident upon the vacuum window, which increases window life.

(6) A series of measurements was made on a QK-338 magnetron to determine its loaded Q as a function of frequency. The intent was to replace the existing waveguide transformer, between the tube output iris and the window, with a filter. To do this, it is necessary that the pass-band image transfer function, and the impedances of the filter and the transformer be as nearly alike as practical. A corruguated waveguide filter was used. This filter is shown on figure 3-29; its characteristics are plotted on figure 3-30. Without the filter, the loaded Q (Q_L) was 132 at the tube resonant frequency of 2789 mc; the VSWR looking into the tube was 17:1. Upon inserting the filter and a variable length insert (to change the insertion phase angle), the Q_L and VSWR were recorded as the filter length was changed (fig. 3-31). From table 3-4 it can be seen that the filter was nearly the optimum length without the insert.



IN1212-217

Figure 3-29. Waffle-Iron Waveguide Filter



IN1212-218

Figure 3-30. Characteristics of Waffle-Iron Waveguide Filter

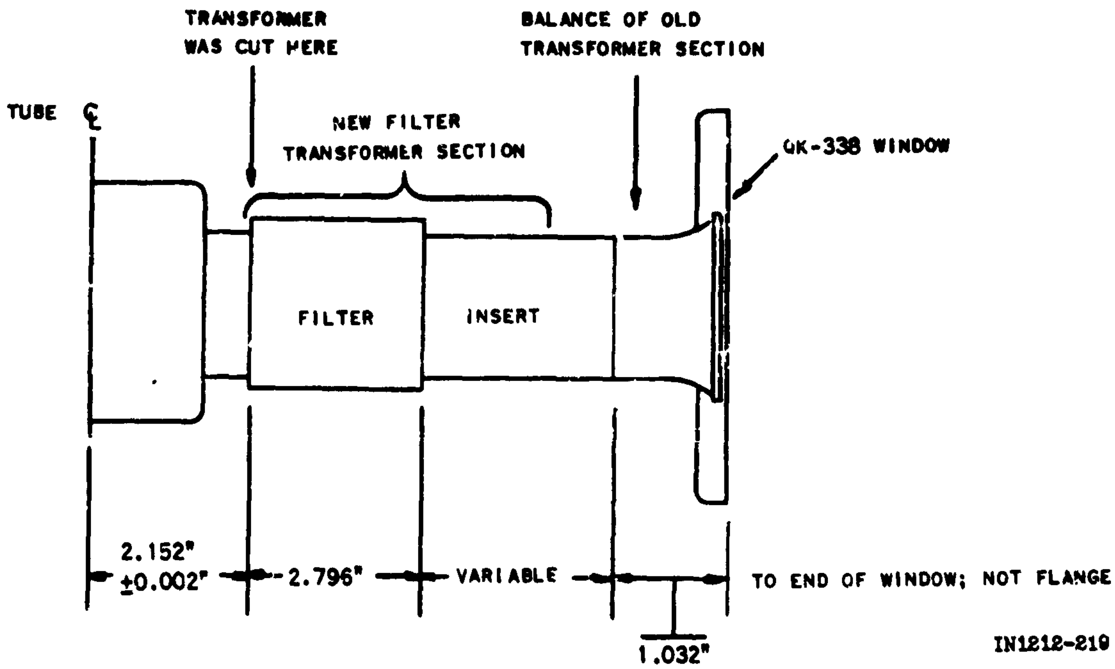


Figure 3-31. QK-338 Magnetron Modified with Harmonic Filter Transformer Section

TABLE 3-4. SUMMARY OF RESULTS ON QK-338 MAGNETRON AND FILTER

Item under Test	F_o	Q_L	VSWR
Tube only, QK-338	2789 mc	132	17
Filter and 1.089-inch insert	2717 mc	20	1.5
Filter and 0.128-inch insert	2766 mc	150	8
Filter and 0.089-inch insert	2772 mc	145	9
Filter and no insert	2794 mc	125	14

h. Summary. Magnetrons generate three types of unwanted signals: moding oscillations, spurious frequency oscillations, and harmonic oscillations. Moding oscillation is minimized by proper matching of the magnetron video impedance to that of the modulator, and proper design of the magnetron strapping. When the spurious oscillations in a particular tube type are identified and their cause is determined, they can be reduced to a negligible level by proper design of the magnetron. Klystrons and traveling-wave tubes require careful matching of the impedances between the tube and power supply to minimize undesired signals. Careful control of the modulation pulse envelope (the rise time should be fast - much faster than that of a magnetron) can reduce splatter generation. In general, a filter that is an integral part of a magnetron can be designed to reduce the harmonic frequency generation by an additional 30 db below the level without the filter. This filter could replace the existing transformer section with only a negligible enlargement of the tube.

3-5. Semiconductor Devices

a. General. Small size, low weight and volume, and elimination of the temperature rise associated with vacuum tubes are among the principal reasons for the increased use of diodes, transistors, and other semiconductor devices.

b. Thermistors.

(1) Positive temperature-coefficient thermistors can be used for arc interference reduction, and to protect semiconductors against high-voltage transients. This ability stems from the sharp change in resistance of the thermistor within a relatively small temperature range. It is limited to relatively slow-cycling applications because of the time required for heating and cooling.

(2) The characteristics of a typical positive temperature-coefficient thermistor are given on figure 3-32. A simple application for arc interference reduction in a conventional switching circuit is shown on figure 3-33. If the thermistor were not in this circuit, the full 120 volts would appear across the switch contacts at the instant of opening. The resultant high-voltage gradient could cause arcing. If the load were inductive, the voltage across the contacts would be increased by L (di/dt) so that:

$$V_c = V_s + L\left(\frac{di}{dt}\right) \qquad (3\text{-}22)$$

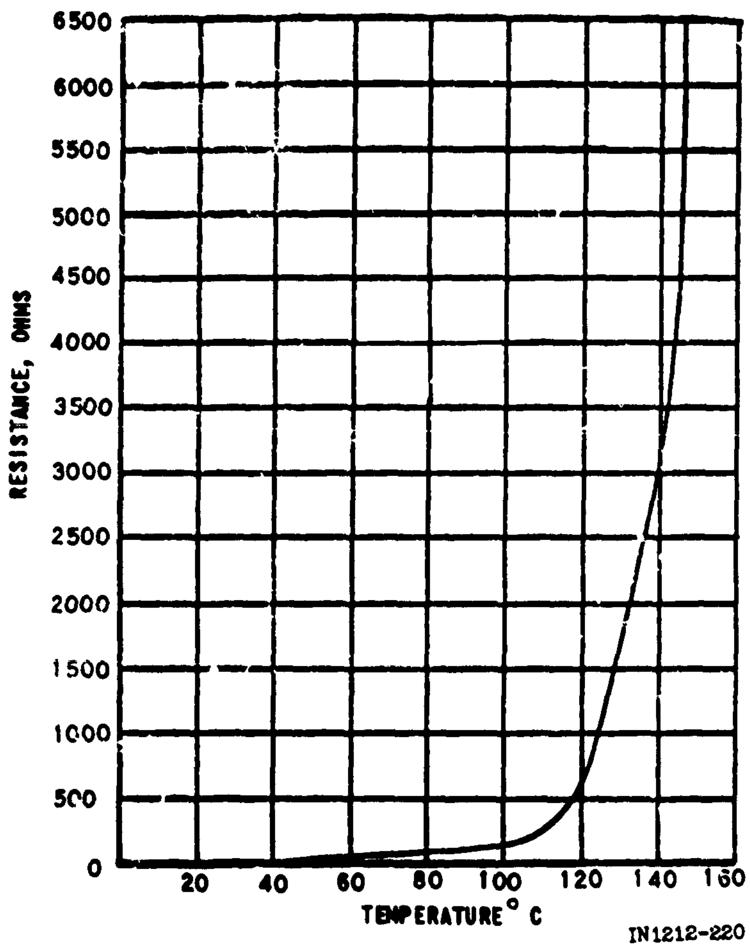


Figure 3-32. Thermistor Characteristics, Positive Temperature Coefficient Type 802-5

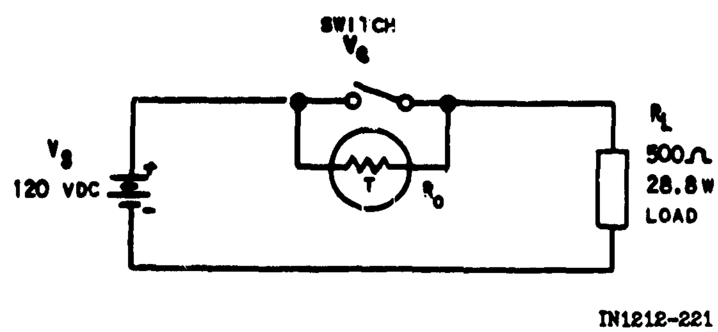


Figure 3-33. Arc Interference Reduction Using a Thermistor

where: V_c = the voltage across the contacts

V_s = the source voltage

(3) In the circuit on figure 3-33, the thermistor dissipates negligible power when the switch is closed. If the ambient temperature is 25°C and the thermistor has stabilized at this value, its resistance (fig. 3-32) would be about 50 ohms. If the switch is then opened to a purely resistive load, the voltage across its contacts would be:

$$V_c = IR_o = \left(\frac{V_s}{R_o + R_L}\right) R_o = \left(\frac{120}{50 + 500}\right) 50 \tag{3-23}$$

$$= 0.218 \times 50 = 10.9 \text{ volts}$$

where R_o = thermistor resistance, and R_L = load resistance. This initial low voltage would prevent arcing or reduce it to a negligible level.

(4) Figure 3-34 shows a circuit in which the switching is done by driving a transistor from the saturated to cutoff state. A thermistor is used to reduce voltage transients between the emitter and collector, and to permit a slow transfer from the saturation state to the cutoff state without excess heating.

(5) Figure 3-35 illustrates how the 1-watt thermistor-protected transistor of figure 3-34 compares to a 10-watt transistor without a thermistor. The arrows indicate the transfer direction from saturation to cutoff. The area under each power curve represents the safe operating range for transistors of that particular wattage rating. The diagram shows that a 1-watt thermistor-protected transistor can be used instead of a 10-watt unprotected transistor. This diagram is applicable to transistors with ratings of V_{CE} and V_{CB} = 120 volts, I_C = 250 ma, and P_C = 1 watt or 10 watts, for the protected and unprotected cases, respectively.

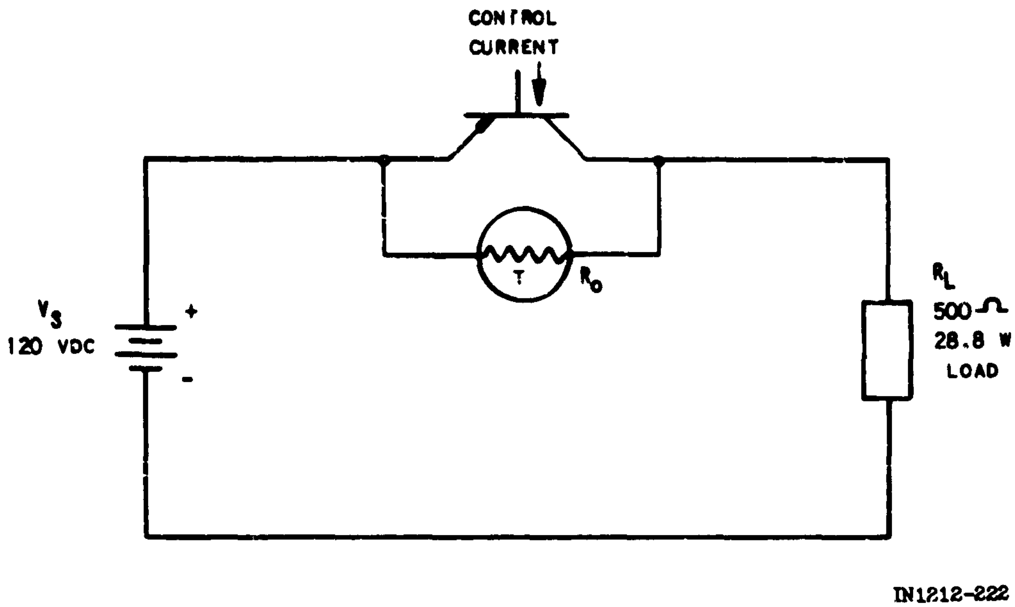


Figure 3-34. Transistor Voltage Suppression Using a Thermistor

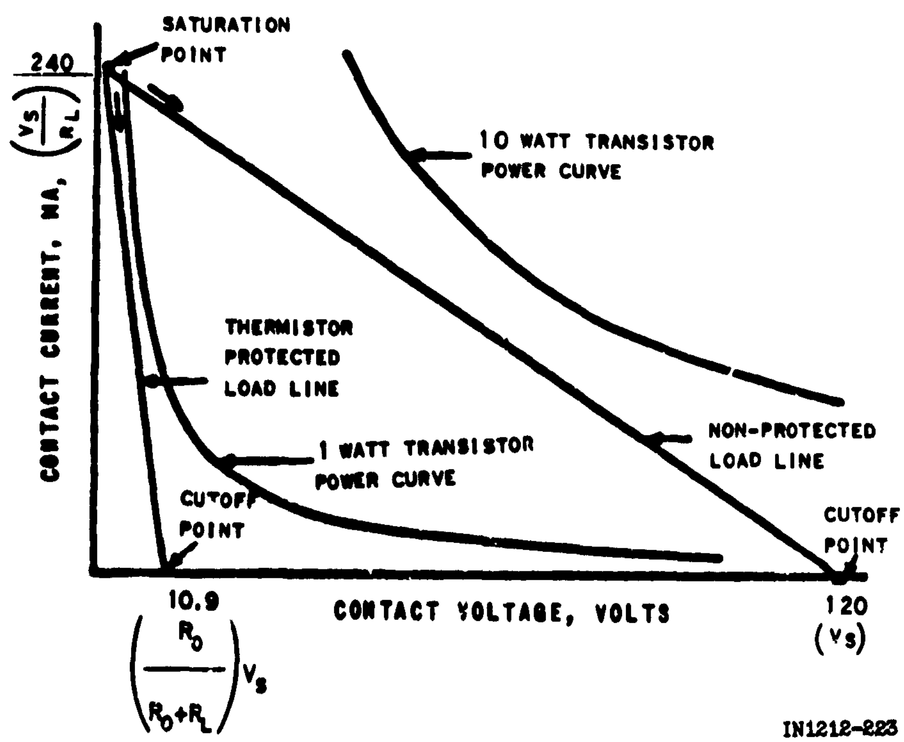


Figure 3-35. Transistor Switching Characteristics

(6) If the load were also inductive (rather than purely resistive), at the instant of circuit opening, the current could not change immediately, and the original current would continue to flow through the load and through the thermistor. Again, with 500-ohms resistance and an unspecified inductance, this original current is:

$$I_o = V_s/R_L = 120/500 = 0.24 \text{ amps} \qquad (3\text{-}24)$$

The voltage across the switch contacts (or semiconductor terminals) at the instant of circuit opening is:

$$V_c = I_o R_o = 0.24 \times 50 = 12 \text{ volts} \qquad (3\text{-}25)$$

Thus, the value of inductance does not appreciably affect the opening voltage, only its duration.

c. Transistors.

(1) Noise. In addition to power at the signal frequency, the output of a transistor stage contains spurious power at all other frequencies. This spurious power is noise. A portion of this noise is introduced into the input stage from the source. Other noise arises from drift and diffusion currents through both the pn junctions and the bulk of the semiconductor. Noise, or random current fluctuations, arise due to the particle, or granular, nature of the current flow. Most transistor noise is attributed to:

(a) Diffusion fluctuations of the minority carriers after crossing a junction

(b) Recombination fluctuations in the base region

(c) Thermal (Johnson) noise in the base resistance

(2) Transistor noise figure. The noise figure of a transistor is defined as:

$$F = \frac{S_i/N_i}{S_o/N_o} \qquad (3\text{-}26)$$

where S_i/N_i is the signal-to-noise power ratio at the input, and S_o/N_o is the signal-to-noise power ratio at the output. For n cascaded transistor stages, the noise figure becomes:

$$F_T = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \cdot G_2} + \ldots . + \frac{F_n - 1}{G_1 \cdot G_2 \cdots G_n} \qquad (3\text{-}27)$$

where: F_T = total overall noise figure

F_n = noise figure of the nth stage

G_n = available gain of the nth stage

As can be seen from equation 3-27, the gain and noise figure of the first stage in a cascaded amplifier largely determine the overall noise figure.

(3) Transistor noise model. The small-signal "T" equivalent circuit of the transistor (figure 3-36) shows the three major sources of transistor noise. This transistor noise model is valid for frequencies less than f_α (the α-cutoff frequency) and greater than 1000 cycles. The mathematical representations of these three noise sources are:

$$\overline{i_e^2} = 2e\, I_e\, B \qquad (3\text{-}28)$$

where: $\overline{i_e^2}$ = emitter noise source

I_e = dc emitter current

e = charge on one electron

B = effective noise bandwidth of the system

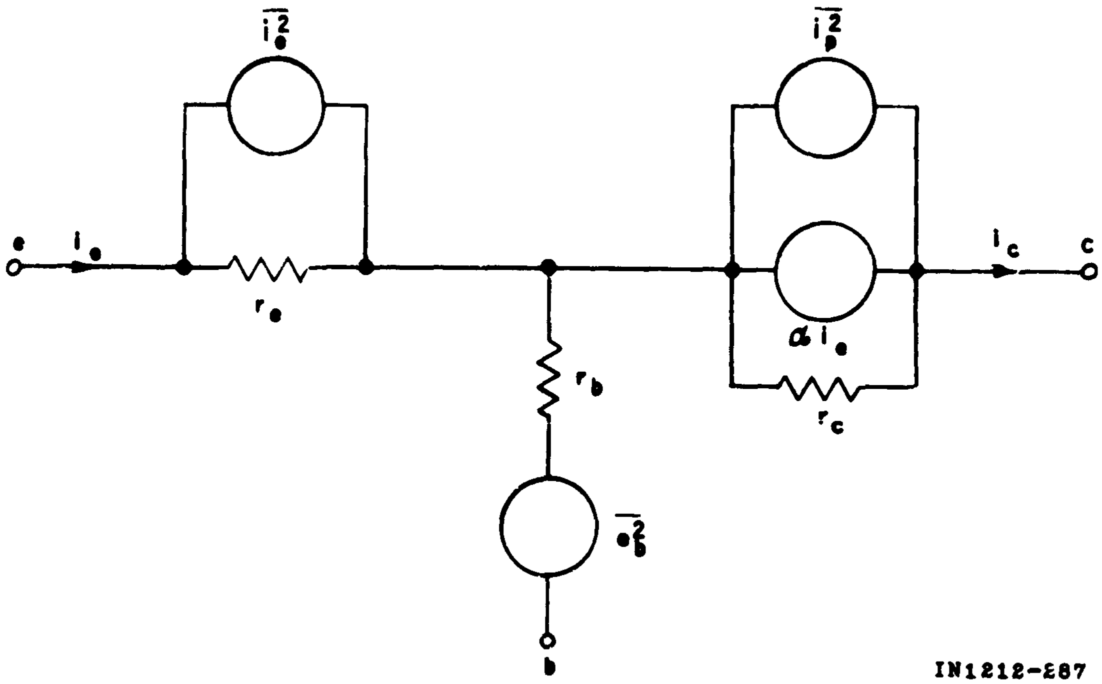


Figure 3-36. Transistor Noise Model, 1KC < f < f_α

$$\overline{i_p^2} = 2e\, I_c \left(1 - \frac{|\alpha|^2}{\alpha_o}\right) B \qquad (3\text{-}29)$$

where: $\overline{i_p^2}$ = noise term in the collector circuit (due to recombination fluctuations in the base) giving rise to additional noise fluctuations in the collector circuit

I_c = dc collector current

α_o = short-circuit current gain at low frequencies

$$\alpha = \frac{\alpha_o}{1 + jf/f_\alpha}$$

Equation 3-29 is valid when $I_{co} \ll I_c\,(1 - \alpha_o)$

where: I_{co} = collector current with zero emitter current

$$\overline{e_b^2} = 4kTr_bB \qquad (3\text{-}30)$$

where: $\overline{e_b^2}$ = thermal noise term due to base spreading resistance r_b

k = Boltzman's constant

T = temperature (°K)

Experimental results are in good agreement with the three-major-source equivalent transistor noise model. Actual results show a slightly higher noise figure which indicates that there are other noise sources in addition to those considered here, but that their contribution is small.

(4) Variation of transistor noise with circuit parameters. The noise figure is a function of (a) frequency, (b) source resistance, (c) emitter current and (d) collector voltage.

(a) Noise figure as a function of frequency. A typical experimental noise figure curve as a function of frequency is shown on figure 3-37. At low frequencies (typically below 1 kc) the noise figure, F, rises at a rate of 3 db/octave, or functionally as 1/f. This type of noise differs from shot and thermal noise which have a flat frequency spectrum. Originally, it was the predominant noise term in transistors, and arose from inhomogeneities in manufacture. More refined methods of design and manufacture have reduced this effect to very low frequencies so that its contribution to the overall noise generation can be assumed negligible over the ordinary range of transistor operation.

At frequencies above 1 kc, the noise figure of the transistor can be accounted for by the incremental model incorporating the three sources of noise previously mentioned. This noise figure may range from 2 to 6 db and is constant up to a frequency of $f_\alpha \sqrt{1 - \alpha_o}$ (f_α is the α-cutoff frequency, α_o is the short-circuit current gain). Above this frequency, the noise figure rises at 6 db/octave, which is accounted for in the noise model by following the usual transistor analysis of letting α vary as $\alpha_o/(1 + jf/f_\alpha)$.

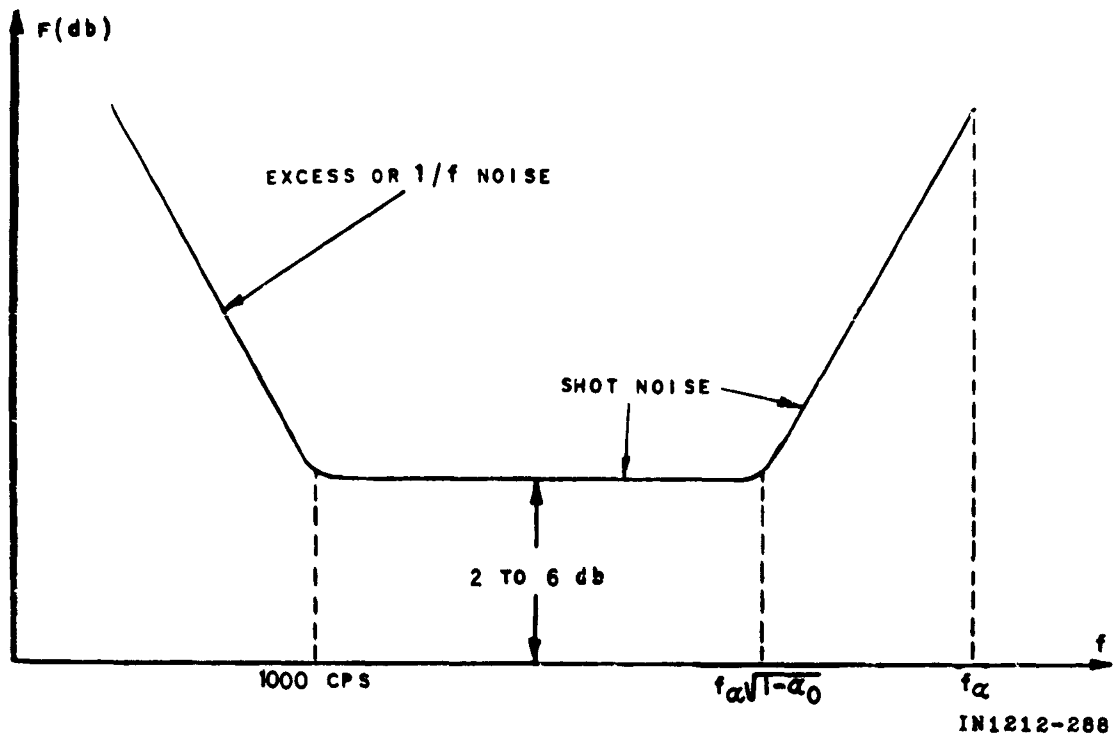


Figure 3-37. Transistor Noise Figure vs Frequency

(b) Noise figure as a function of source generator resistance.

The noise figure is a function of the source generator resistance, and possesses an optimum value. The typical variation of noise figure, F, with source resistance, R_s is shown on figure 3-38. The optimum value of R_s for the common base configuration is given by:

$$R_{s,\,opt} = \left(r_b^2 + \frac{2r_b r_e}{1 - \alpha_o}\right)^{1/2} \qquad \begin{array}{l} r_e \ll r_b \\ f < f_\alpha\sqrt{1 - \alpha_o} \end{array} \qquad (3\text{-}31)$$

where r_e is the emitter resistance and the other symbols have the same meanings as in the previous discussions.

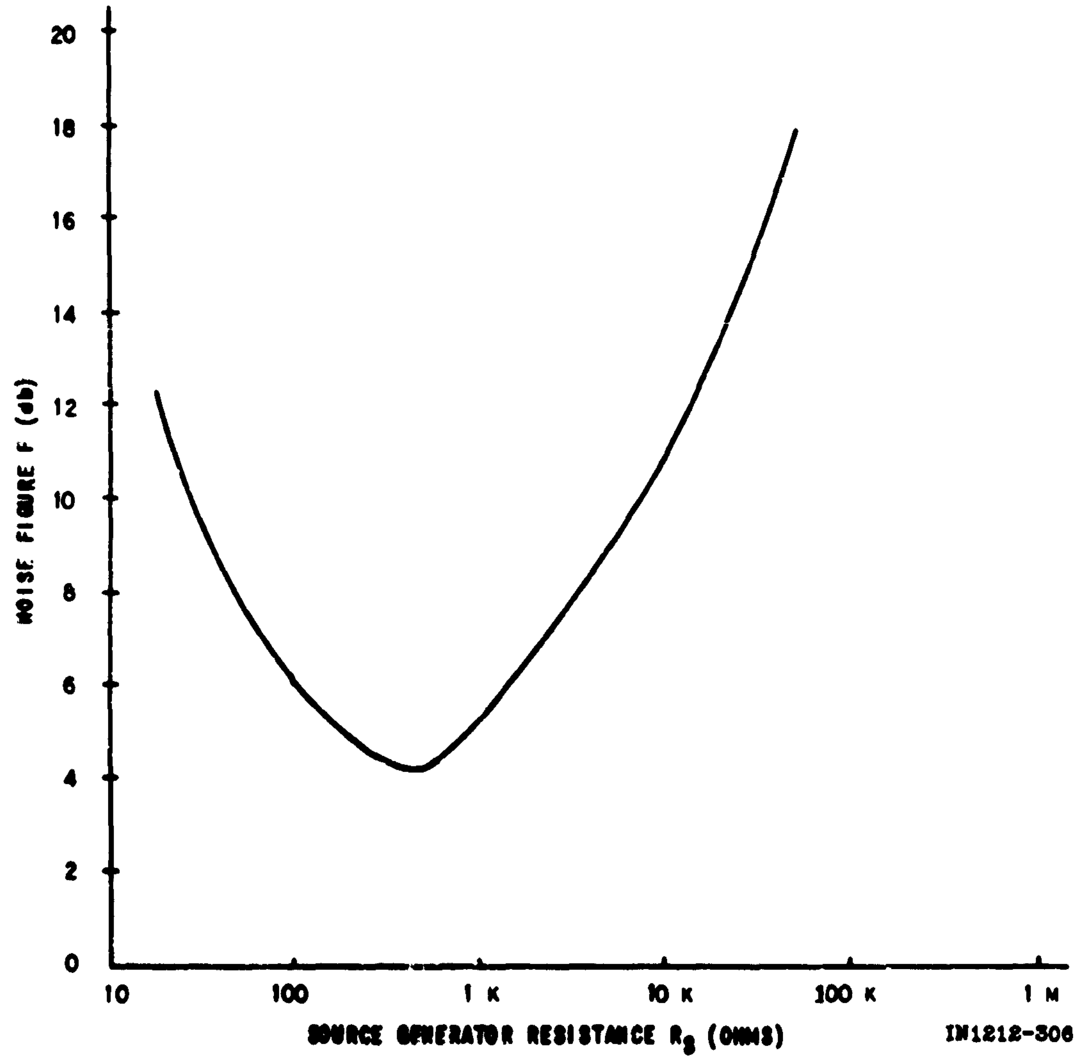


Figure 3-38. Typical Dependence of Transistor Noise Figure on Source Generator Resistance at Fixed Frequency and Operating Point

This optimum value of R_s is not critical as shown by the curve of figure 3-38. A typical noise figure may change by only 0.5 db for a variation in R_s of from one-half to twice the optimum value; and may change by only 2.5 db over a range from 1/5 to 5 times the optimum value.

(c) Noise figure as a function of emitter current. To keep the noise figure as low as possible, the emitter current, I_e, must be kept as low as possible. The lowest practical value for I_e is typically of the order of 0.3 ma. When I_e is smaller than this value, short-circuit current gain (α_o) begins to decrease. The α_o should be as close to unity as possible in order to decrease recombination fluctuations. Figure 3-39 shows a typical example of the relationship between emitter current and transistor noise figure.

(d) Noise figure as a function of collector voltage. A high value of collector voltage increases the transistor noise figure. As shown on figure 3-40, the noise figure of a typical transistor remains constant as the collector voltage increases until, at about 10 volts, it begins to rise sharply.

(e) Other factors influencing the noise figure. The noise figure is approximately the same for all three transistor configurations (common-base, common-emitter, common-collector). For minimum noise figure, α should be close to unity, f_α should be high, and r_b should be small.

(5) Interference. When analyzing interference in transistor circuits, it is convenient, as with vacuum-tube circuits, to evaluate it in terms of small-signal and large-signal responses. Under small-signal operation, it is relatively safe to assume linear operation. Interference responses under these conditions are common in communications-type

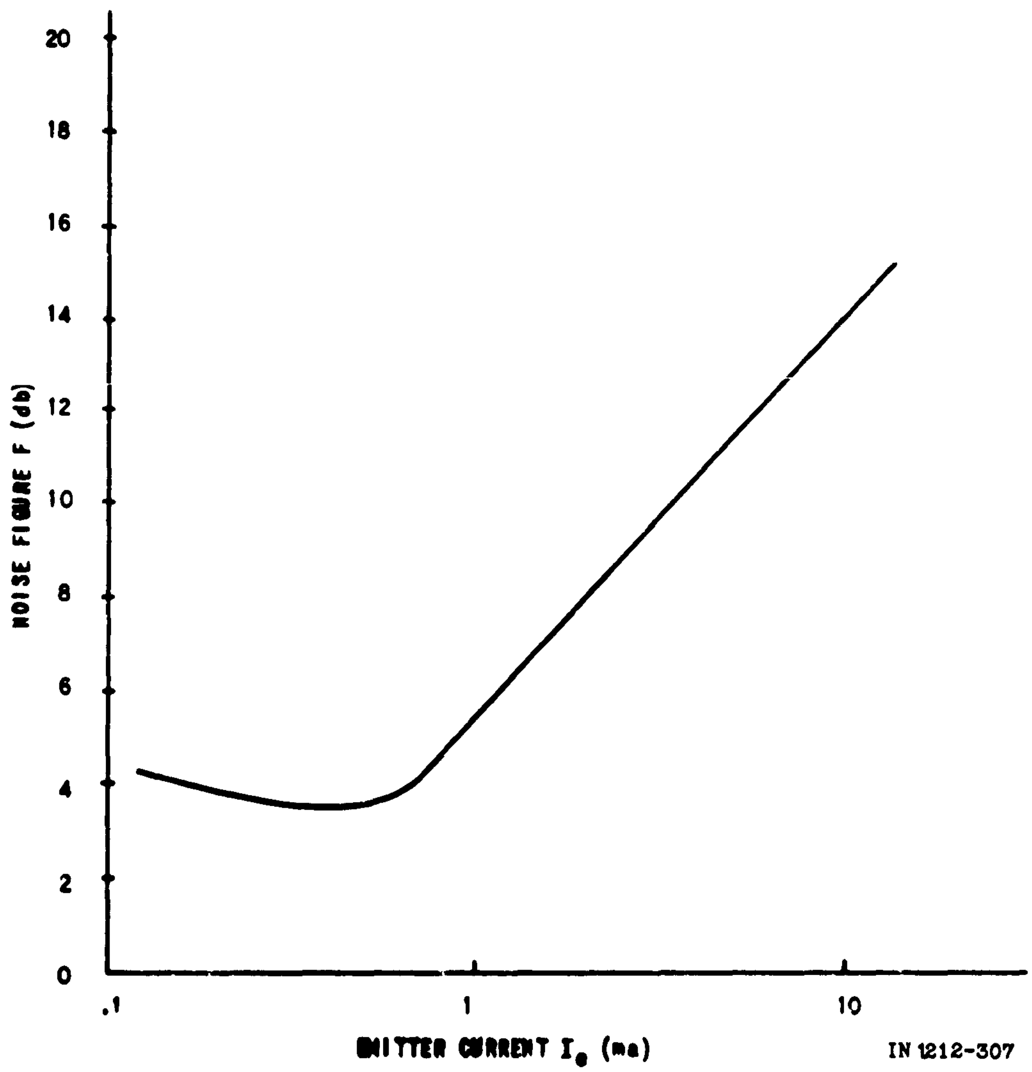


Figure 3-39. Typical Dependence of Transistor Noise Figure on Emitter Current

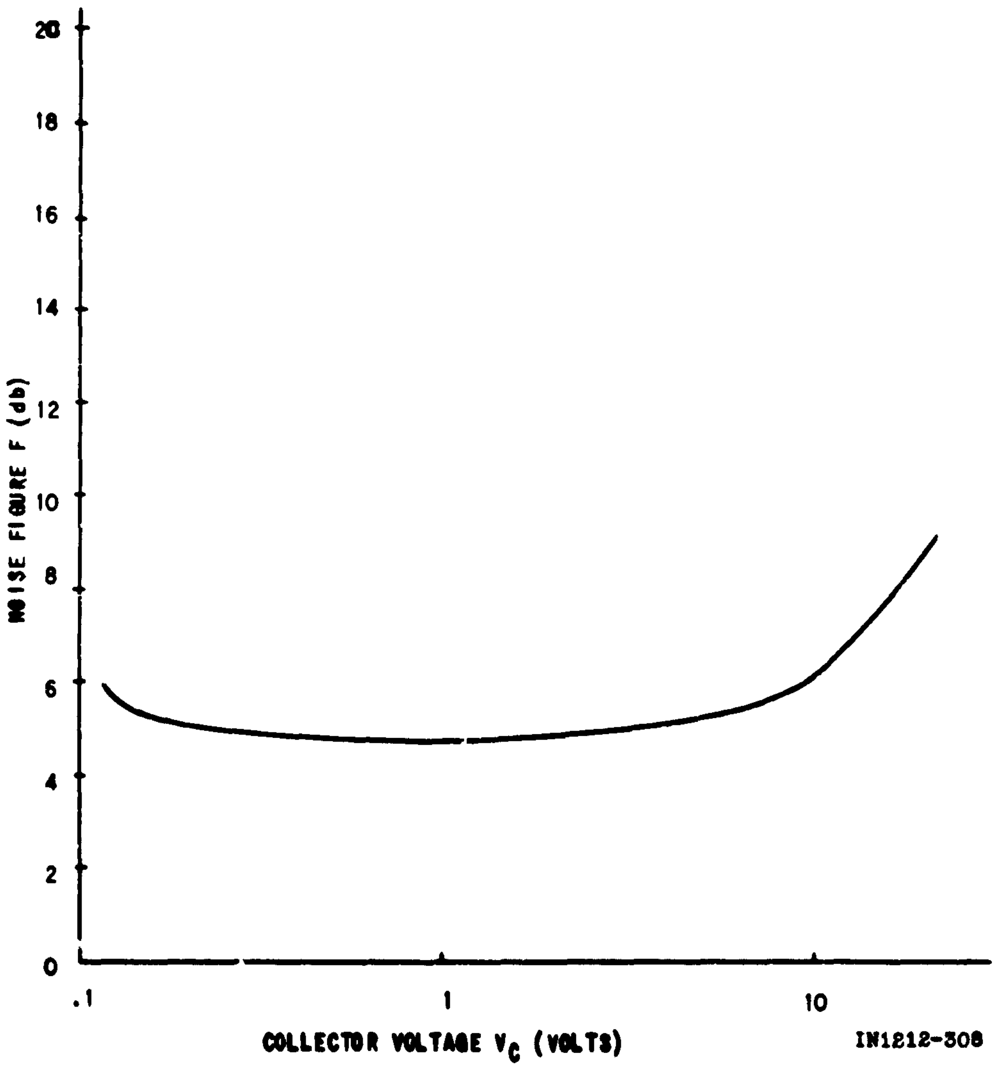


Figure 3-40. Typical Dependence of Transistor Noise Figure on Collector Voltage

circuits. They include images, spurious, harmonic and if responses. Signals that are large compared to the normal signal levels for which communication circuits are designed frequently drive these circuits into non-linear regions. Such non-linear operation gives rise to spurious interference responses that would not otherwise occur. Interference responses such as cross-modulation, intermodulation, desensitization, and blocking occur under non-linear operation.

(6) Cross-Modulation.

(a) Non-linear characteristics of transistors may cause cross-modulation in high-frequency amplifiers exactly as do the non-linear characteristics of vacuum tubes. There is a marked difference in behavior of alloy junction and drift transistors with respect to cross-modulation. At frequencies below about 1 mc, this difference in behavior is caused by the difference in base resistance. At higher frequencies, the diffusion capacitance must be considered, but the base resistance still remains important.

(b) The cross-modulation characteristic in transistors, as in vacuum tubes, can be described by the rms value of an interfering signal, V_e, with modulation index, m, that produces a cross-modulation index of 1 per cent. For transistors, it can be shown that:

$$V_e = \frac{0.1(V_T + I_B R_B)^2}{mV_T\sqrt{V_T - 2 I_B R_B}} \qquad (3\text{-}32)$$

where $V_T = kT/e$ (about 26 mv at 23°C), k is the Boltzmann constant (1.380×10^{-23} joule/deg K), T the temperature in °K, e the energy associated with 1 electron volt (1.602×10^{-19} joule), and I_B is the base current in amperes. In this equation, the value of R_B must include any significant output resistance of the source. Curves showing typical values of V_e as a function of the base current are plotted in figure 3-41. Of particular interest is that value of base current at which V_e is infinite: $(I_B = V_T/2R_B)$.

(c) For drift transistors, the value of base current that causes V_e to be infinite is approximately one decade higher than the value for the alloy junction transistor. The discussion can be related to collector current by using the current gain factor, $\beta = I_E/I_B$. On figure 3-42 and 3-43, the value V_e (100 per-cent modulated) necessary to produce one per-cent cross-modulation index is shown as a function of base resistance with β as a parameter.

d. Diodes. Semiconductor diodes can generate broad-band interference over a wide frequency range. A diode is incapable of instantaneously switching from the conducting state to the nonconducting stage. When a diode, that has been passing a forward current, is suddenly biased in the reversed direction, current carriers in transit through the semiconductor material are trapped and must suddenly reverse their direction of flow in the circuit. This occurs at the time and point that the applied voltage goes through the zero reference. The result is a sharp surge of reverse current through the diode and the load. This process of switching, or reversing, results in the generation of interference. In general, the magnitude and frequency spectra of this interference are functions of both the diode's transient characteristics and the load.

(1) Interference in descending amounts is produced by silicon diodes, selenium rectifiers, and vacuum tubes. Similarly, in descending order, interference is produced in frequency conversion and by inverters, rectifiers, and dc transformers.

(2) There are no specific rules for the reduction of diode interference. In general, diode interference can be greatly minimized or avoided by careful diode selection in accordance with the following rules:

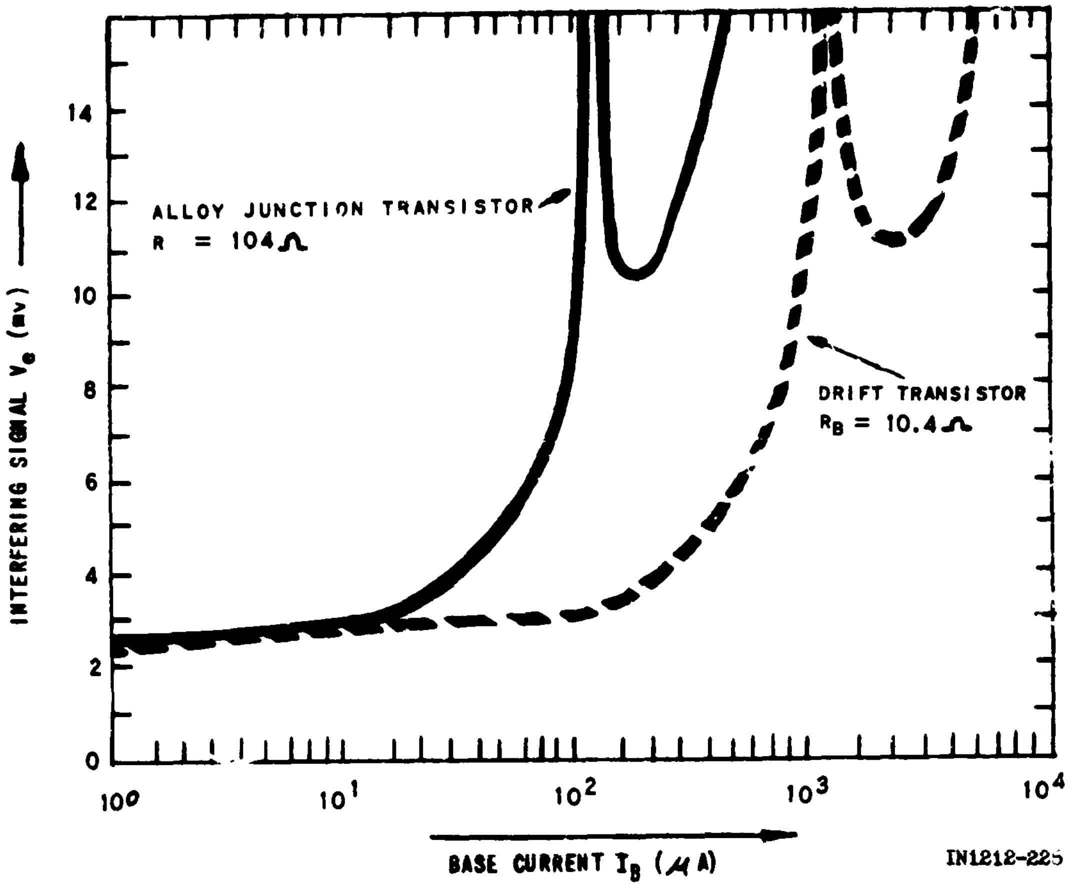


Figure 3-41. 100% Modulated Interfering Signal Necessary to Produce a Cross Modulation Index of 1%

(a) Select a diode that will operate at the lowest current density in proportion to the manufacturer's maximum current rating

(b) Select a diode with the highest working and peak inverse voltage ratings, so that they will never be exceeded

(c) Use the lowest possible diode switching rate. In most applications, the severity of interference is a direct function of the switching rate: the faster the diode switches, the worse the interference becomes

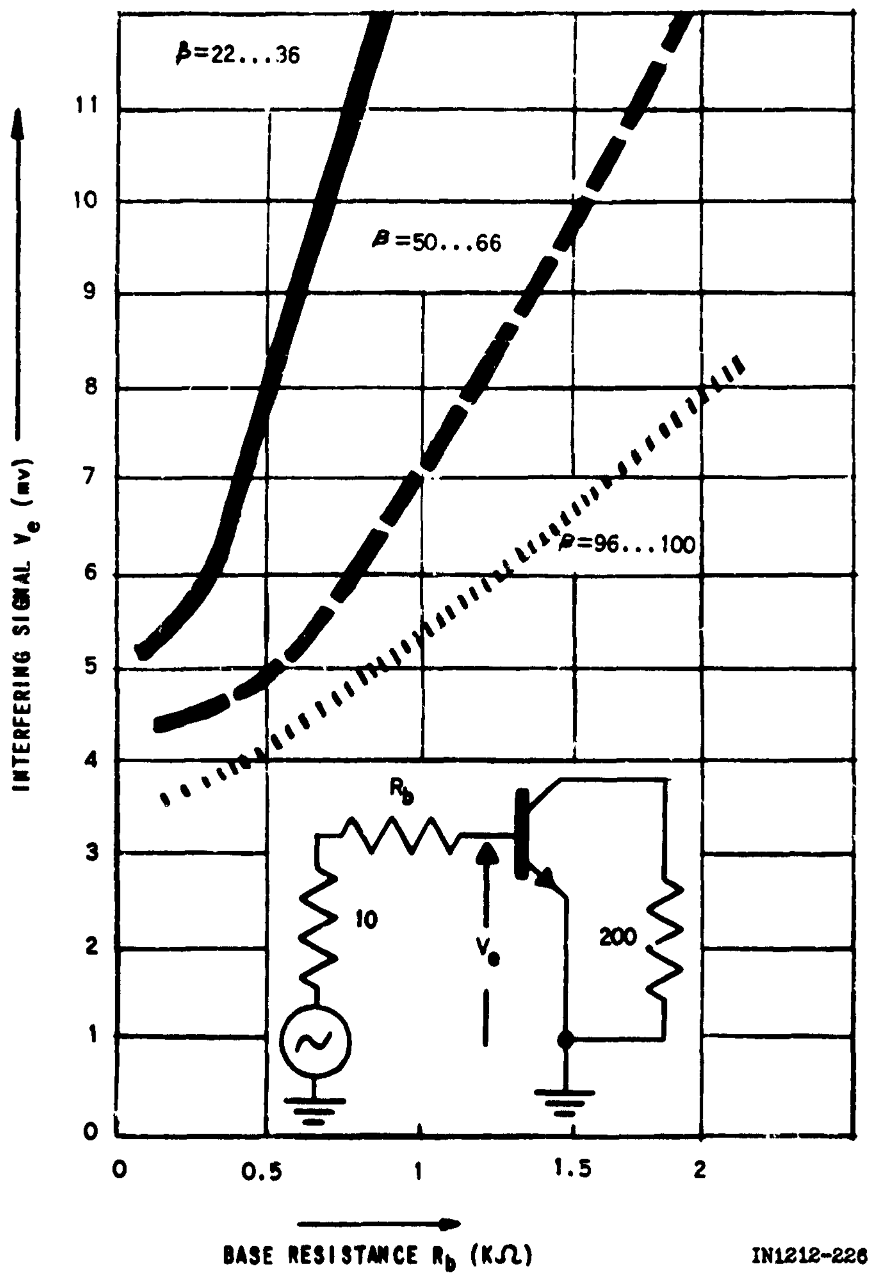


Figure 3-42. 100% Modulated Interfering Voltage Necessary to Produce 1% Cross Modulation Index at 1 kc for an Alloy Junction Transistor

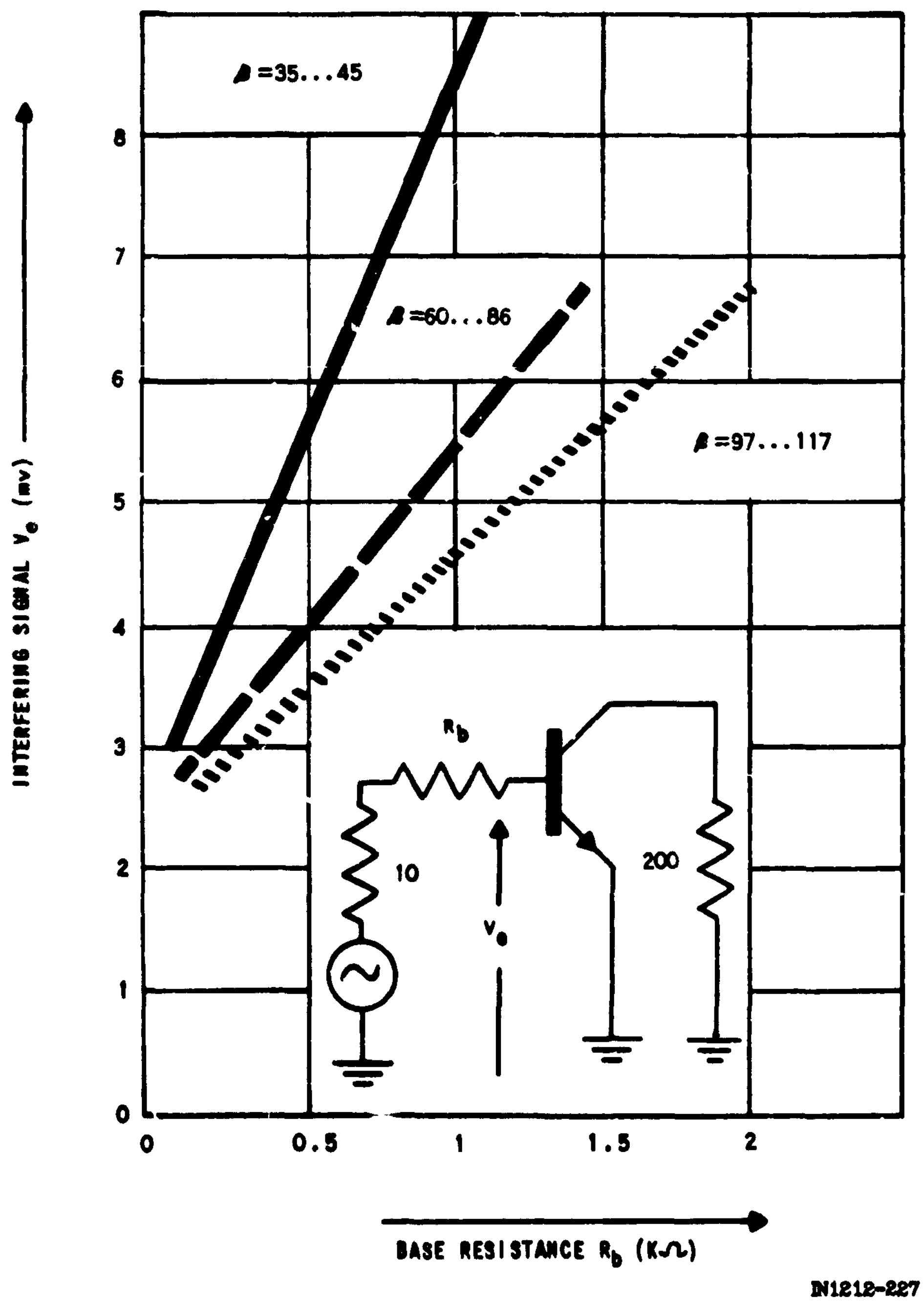


Figure 3-43. 100% Modulated Interfering Voltage Necessary to Produce 1% Modulation Index Measured at 1 kc for a Drift Transistor

(3) Because of the number and variety of diode types and their applications, it is often advantageous to determine by interference tests which are the best diodes available for a particular application. Such tests can save considerable cost in time and equipment and can result in reduced size and weight. When reducing diode-generated interference, any filter components or networks that are used will have maximum effectiveness if they are physically applied directly at the diode. In many cases, a complete unit is designed by packaging the diodes and associated filter networks in a single metallic container. This method is particularly important in applications where space is limited. Such practice lends itself most readily to designs where small current diodes are involved. The single-unit (diodes plus filters) design prevents the diodes and their wiring from conducting or radiating rf voltages to adjacent circuits.

(4) Typical interference reduction measures are shown in figures 3-44 and 3-45. For 60-cycle applications, the inductance, L, is usually between 100 and 300 μh; while for 400-cycle applications, it may be between 250 and 750 μh. The circuit in figure 3-44A reduces interference by controlling the transient current; the circuits in figures 3-44B and 3-44C employ filtering. The circuit in figure 3-44C is generally used at higher line frequencies and when greater attenuation is needed than that afforded by inductors alone.

(5) A capacitor across the output of a bridge circuit is effective in reducing both ripple and interference. For 60-cycle applications, a capacitor of from 0.1 to 0.5 μf can be used. When the dc output from the bridge circuit exceeds one amp, a power-line filter configuration should be used for reducing both ripple and interference. A typical circuit is shown on figure 3-45A. The L-type filter is formed by C_1 and L; C_2 represents a high value

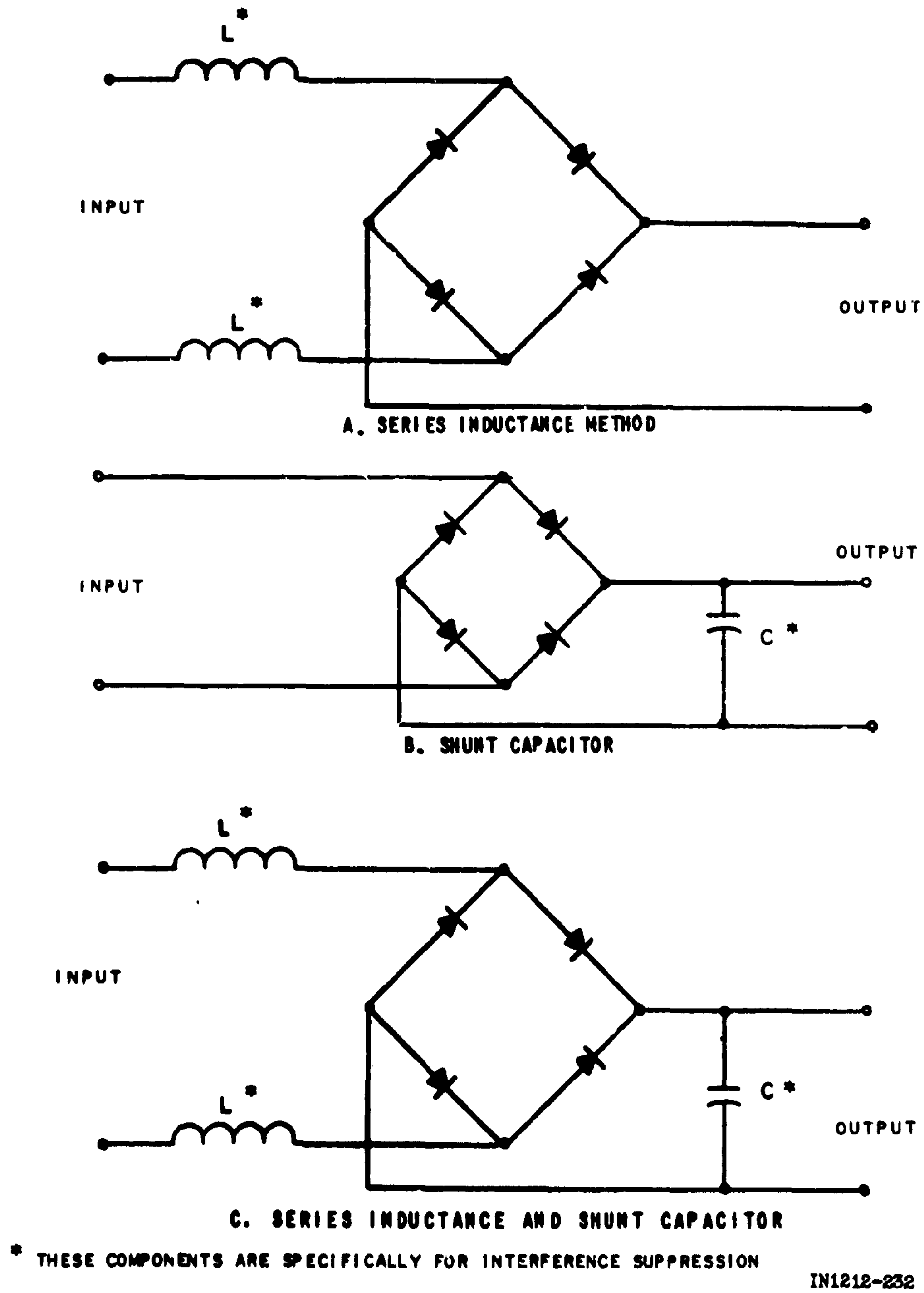


Figure 3-44. Interference Reduction of Diode Rectifier Circuits with Series Inductance and Shunt Capacitor

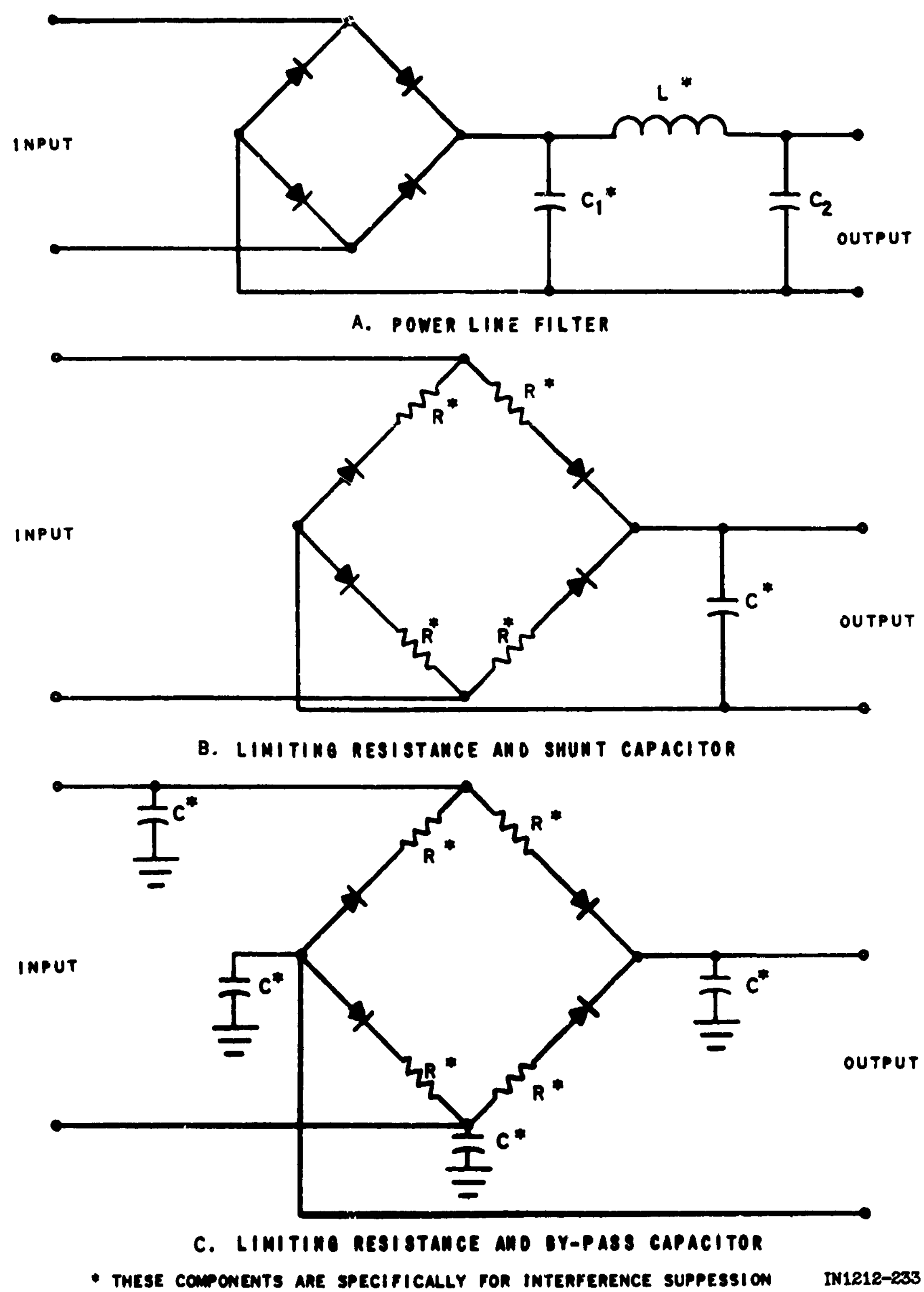


Figure 3-45. Interference Reduction of Diode Rectifier Circuits with Power Line Filter, Limiting Resistor and Capacitors

of capacitance such as that normally used for ripple filtering. The inductor limits surge current to C_2 and the load by acting as a reactance voltage divider. If power and heat dissipation can be tolerated, resistors of the order of 1 to 10 ohms may be inserted in the bridge, as shown on figure 3-45B. These resistors limit the magnitude of switching and rf transients. The capacitors shown on figure 3-45C bypass rf currents to ground. In 60-cycle bridge networks, the resistors alone are usually sufficient for interference control. For 400-cycle applications, capacitors of 0.05 to 0.25 µf will effectively bypass rf currents.

e. Zener Diodes.

(1) If a Zener diode is biased in the forward direction (positive to anode), considerable current will flow when the barrier potential is exceeded (fig. 3-46). When a source of low voltage is applied to the diode in the reverse direction (positive to cathode), the junction back-resistance remains quite high, and junction current is of the order of several µa. As the reverse potential is increased, the junction reaches a critical point, and the diode avalanches. This breakdown is not destructive as long as the diode's dissipation capabilities are not exceeded; the device may be cycled in and out of this region as often as necessary. When avalanche breakdown is reached, the normally high back-resistance drops to a low value; and the junction current increases rapidly, limited principally by circuit resistance. As the voltage is increased beyond the breakdown point, the diode current increases proportionately, but the junction voltage remains essentially constant.

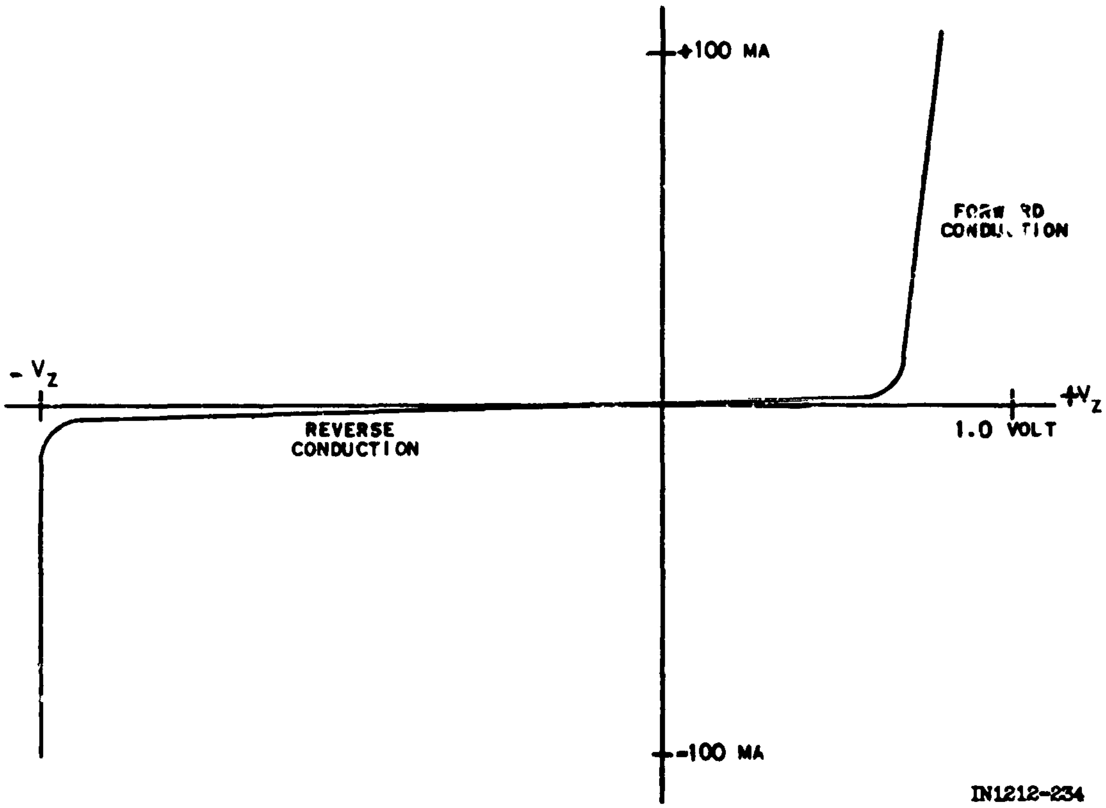


Figure 3-46. Current and Voltage Characteristics for a Typical Zener Diode Regulator

(2) Whenever electric power to inductive devices is controlled by contacts that initiate and interrupt the flow of current, the problem arises of protecting the contacts from erosion and reducing both inductive voltage and interference. These problems are present in both ac and dc circuitry. Particular consideration should be given to transistor circuits. Excessive potentials can destroy the transistor junctions. Zener diodes are used in manual as well as transistor switching circuits for reducing excessive potentials. When a switch in an inductive circuit is opened, the magnetic field in the coil collapses, and a voltage is generated equal to L di/dt (where L is the coil inductance and di/dt is the

time rate of change of the decay current). This voltage is frequently many times larger than the supply voltage and is sufficient to maintain an arc across the opening switch contacts. Repeated arcing causes erosion, pitting, and general physical deterioration of the contacts and results in high-contact resistance, increased maintenance, and most important, generation of interference. To obtain maximum contact protection with optimum circuit operation, it is necessary to design the proper interference reduction circuit (fig. 3-47) and to specify the proper type and rating of the semiconductor device. For all ac circuits, the circuit design of figure 3-47C is necessary since voltages of both polarities are involved. In some applications, a delay in drop-out time of a relay or contactor may be advantageous. The duration of this delay can be controlled by using the circuit of figure 3-47B with a variable resistor.

f. Tunnel Diodes.

(1) The tunnel diode is a heavily doped semiconductor junction that exhibits negative resistance current-voltage characteristics when the diode is biased positively in a range from approximately 50 to 300 mv. The tunnel diode presents some difficulties in use. Its low negative resistance characteristic makes it difficult to provide a dc biasing supply of low enough impedance. This poses a problem in reducing parasitics in the external circuitry. The gain of a tunnel-diode amplifier is also a marked function of load conductance. A change of 1 per-cent in the load conductance produces a 30 per-cent change in resonant gain. Measurements on such circuits should introduce as little additional conductance as possible.

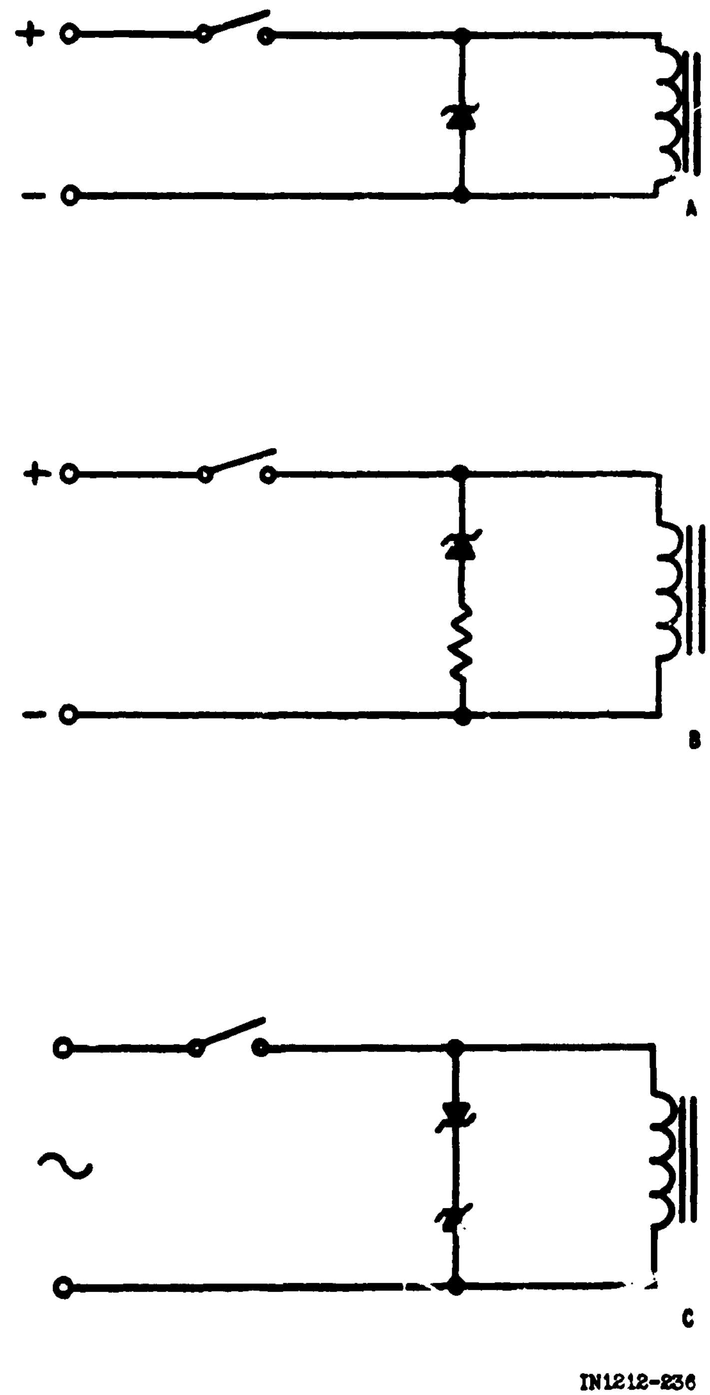


Figure 3-47. Interference Reduction Circuits Utilizing Zener Diodes

(2) Because the tunnel diode has high conductance, it can be operated at kmc frequencies only if the associated circuitry has a low series impedance. Pigtail connections to the device are to be avoided because the lead inductance becomes excessive. A standard crystal cartridge, normally used for silicon diodes, has too much inductance for a tunnel diode.

(3) Some complex interference problems were found in project "Lightening", a 1 kmc tunnel diode computer using discrete miniature component circuitry. The extremely high-speed operation, with semiconductor devices and reduced size of equipment, made the design very difficult. Three main problem areas arose from high-speed operation:

   (a) Delays due to wire length were significant and required optimum location of components

   (b) Signal waveform distortion was greatly increased

   (c) Signal crosstalk became sufficiently large to require shielding of a large part of the wiring. One type of shielding was accomplished by using a channel wiring assembly; the wires, already covered by insulation, were placed in slots. The surfaces of these slots were metal-plated, and the resulting configuration was virtually a coaxial transmission line. The circuits were placed on wafers, and the wafers were placed in a holder (fig. 3-48).

A large part of the interference problem was due to common ground return paths. Designing to isolate the ground return paths as completely as possible reduced the coupling to an acceptable level and permitted the desired operating speed to be approached.

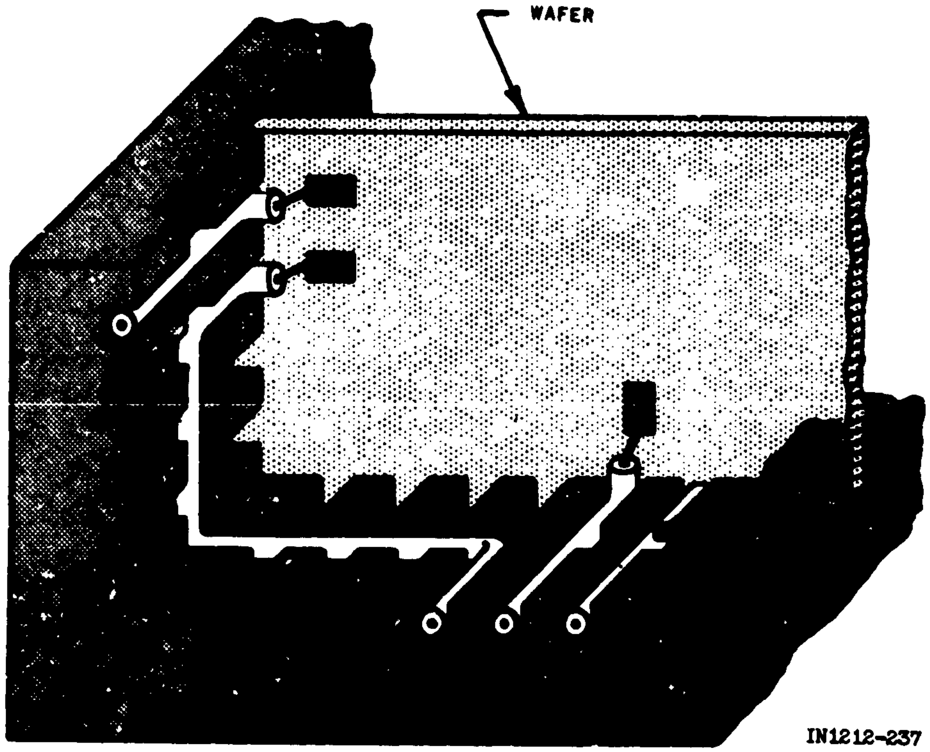


Figure 3-48. Wafer Interconnection with Transmission Lines

g. Silicon-Controlled Rectifiers.

(1) General. The inherently regenerative turning-on action of silicon-controlled rectifiers (SCR's) causes them to switch very rapidly and tends to shock-excite supply lines. Fast rates of rise of current can cause voltages to be induced across the distributed inductances of a supply line. In the presence of distributed line capacitance, this action causes a redistribution of charge on the line. Generally, the redistribution of charge will be oscillatory at a fundamental frequency determined by the parameters of the supply line. For most common types of power distribution lines, the fundamental frequency is usually between 0.250 mc and 2 mc. An SCR may be considered as a generator, producing high-frequency voltages that can give rise to interference, or act upon other SCR circuits. When acted upon

by line disturbances, an SCR circuit may be considered as a receiver that is sensitive to high-frequency voltages. Generally, these voltages act upon the anode of the device directly; or find their way into the trigger circuit and cause the SCR to fire spuriously.

(2) Interference reduction design procedures. There are two possible cases of ground circuitry configuration to consider in SCR operation. In one case an rf ground is not available at the equipment that houses the SCR; for example, in most residential wiring. The other case is where an rf ground is readily available: for example, a military power supply may be self-contained in a metal enclosure mounted in a metallic rack. Such a large expanse of metal can be considered an effective rf ground.

(3) Conducted interference. When switching speed is not critical, the basic approach in designing interference reduction for SCR circuitry is to localize and contain the inital high rate of rise of current to as small a section of the circuit as possible. This can be done by placement of small inductances in the circuit between which this high rate of rise current is to be localized. Generally, a return path must be furnished for the high-frequency component of this current. A small noninductive capacitor will provide this return path. Figure 3-49 shows a good interference reduction design approach for a system that does not have an rf ground accessible. Figure 3-50 shows an approach for the system in which an rf ground is accessible. The values of the components to be used depend upon the amount of interference reduction desired, the nature of the supply line, the layout of the circuit, the nature of the load, and generally, the power level of the circuit. In a conventional 117-volt ac supply line, with a load under 1 kw, the following values have been found to yield good results: L = 60 μh and C = 0.01 μf.

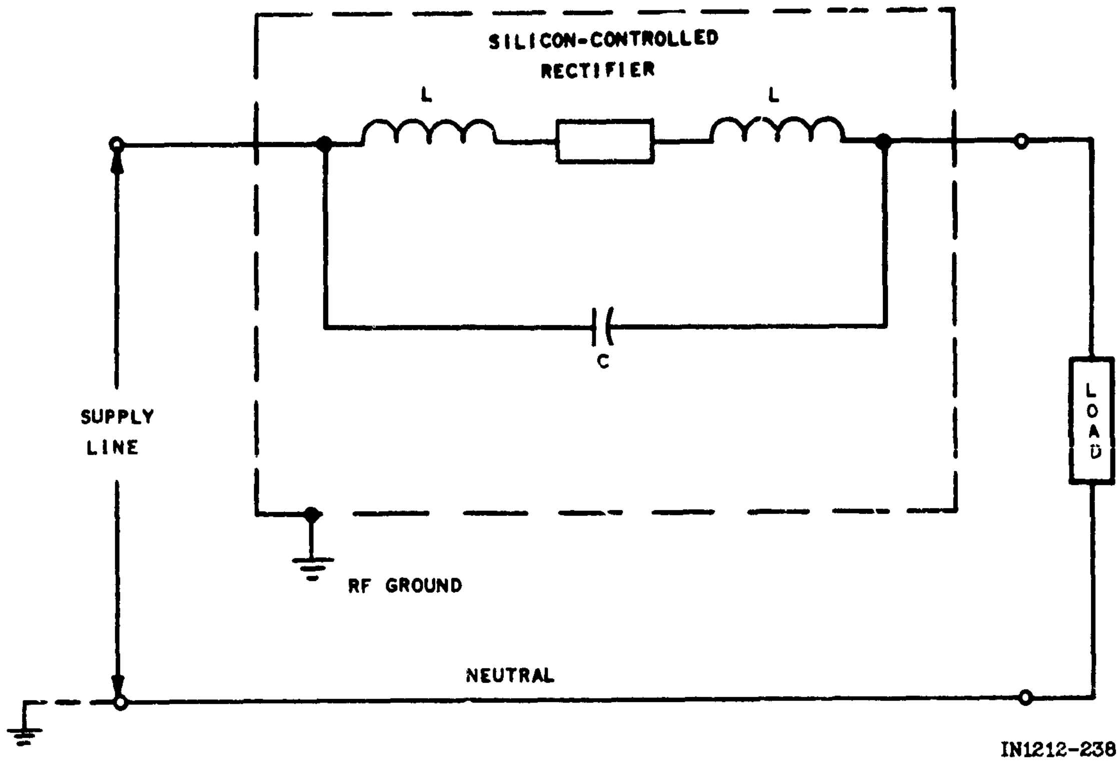


Figure 3-49. Conducted Interference Suppression with RF Ground Inaccessible

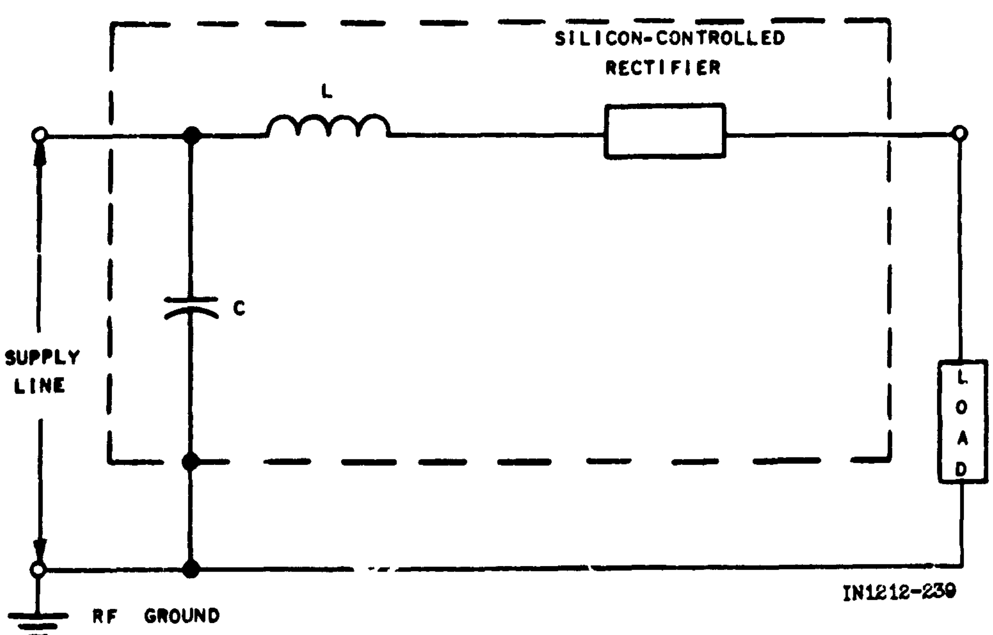


Figure 3-50. Conducted Interference Suppression with RF Ground Accessible

The inductance need be in the circuit for only the few microseconds of the turn-on switching interval. For this reason, the size of the inductance, even in high-capacity circuits, will be small since it is allowed to saturate after the turn-on interval. The nature of operation of the SCR is relatively unimportant in connection with the generation of interference. The same general considerations, in designing for reduced interference, are equally valid whether the SCR is used in a phase-controlled system or in an inverter or chopper system.

(4) Interaction. The SCR system acts as a receiver of any voltage transients generated elsewhere in the circuit. These transients can act on either or both the trigger circuit for the SCR or the anode of the SCR in the main power circuit. Interaction causes the SCR system acted upon to follow or track, completely or partially, another SCR system. In addition, various types of partial turn-on, depending on the nature of the trigger circuits, may occur. The elimination of interaction phenomena must take total circuit layout into consideration. When an SCR circuit is acted upon with its gate circuit disconnected (open or terminated gate), the interaction is usually attributable to the rate of rise of forward voltage. When energizing the circuit, such as by a contact or circuit breaker, applicable rate of rise of forward voltage specifications for the device must be met. Once the circuit is energized, the SCR will sometimes respond to high frequencies superposed on its anode supply voltage. Applicable specifications for the SCR must meet this condition, or steps taken to attenuate the rate of the voltage rise. Because of the nature of anode circuit interaction, it is unlikely that an SCR will track another SCR circuit over the full phase control range. It will usually lock-in over a very limited range near the top of the applied anode voltage half-cycle where its sen-

sitivity to rate of voltage rise is greatest. The best way to reduce this type of interaction is either to increase the capability of the SCR to withstand the rate of voltage rise, or reduce the rate of rise of positive anode voltage. Negative gate bias, a small capacitor (not over 0.1 μf) connected directly across the SCR, resistor-capacitor network elsewhere in the circuit, or a combination of these three measures is recommended. When using capacitors for transient reduction work, only low-inductance capacitors with very short lead lengths should be used.

(5) Radiated interference reduction design. Good electrostatic shielding is the best approach to reduce radiated interference; the high rate of rise of current should be confined to as small a volume as practical. The dashed lines on figures 3-49 and 3-50 indicate the volume of the circuit that must be enclosed within shielding. The shielding is then brought to a solid rf ground.

(6) Interference on trigger circuits. The unijunction transistor (UJT) is an ideal device for use in firing circuits of SCR's because of its stable firing voltage, low firing current, and wide operating temperature range. SCR firing circuits using UJT's are simple and compact, with low power consumption and high effective power gain in phase-control circuits. In addition to good design practice in the entire SCR circuit configuration, as well as in the triggering circuits, the following steps for interference decoupling can be taken when UJT's are employed in the SCR's trigger circuit. There are two major types of interaction that can act on a trigger circuit: interaction directly from a supply line and interaction from an SCR gate circuit. Both of these may cause the trigger circuit to fire prematurely and give rise either to spurious firing or complete or partial tracking of the

SCR's in the circuit. The response of the trigger circuit to incoming transients determines the degree of interaction.

(a) <u>Decoupling the UJT circuit from supply transients.</u> Either one or a combination of the following suppression devices will give effective interference decoupling against line voltage transients acting on the UJT firing circuit:

1) Control (isolation) transformer with an rf filter across its secondary

2) Boot-strap capacitor between base two and the emitter of the UJT

3) Thyrector diode across the supply to the unijunction circuit

The value of the boot-strap capacitor C_1 (fig. 3-51A) should be chosen so that the voltage divider ratio of C_1 and C_2 is approximately equal to n, the intrinsic standoff ratio of the UJT:

$$n = \frac{C_1}{C_1 + C_2} \qquad (3\text{-}33)$$

If this condition is met, neither positive nor negative transients on the unijunction supply voltage will fire the UJT.

(b) <u>Decoupling UJT circuits against SCR gate transients.</u> Negative voltage transients, appearing between the gate and cathode of the SCR's, can cause erratic firing when transmitted to the UJT. When transformer coupling is used, these transients can be eliminated by using a diode bridge in the gate circuit of the SCR (fig. 3-51B). Negative transients often arise in SCR gate circuits, in impulse-commutated circuits, and, under certain conditions, in ac phase-control circuits.

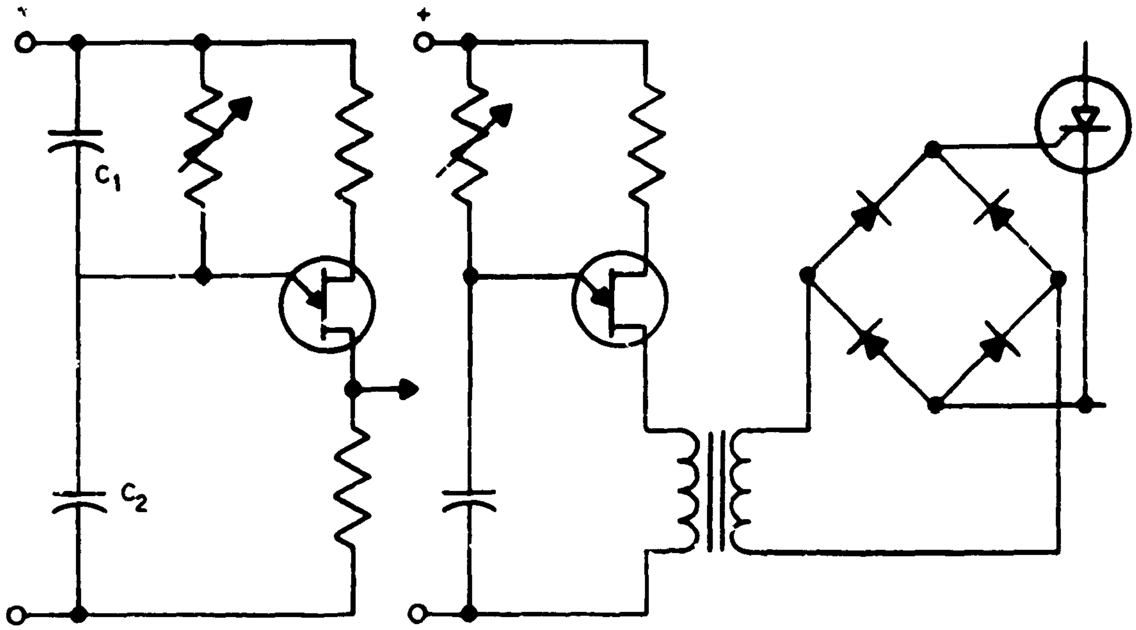


Figure 3-51. Circuits for Decoupling of Unijunction Transistor from Voltage Transients

(7) Good design practices to minimize sources of SCR interaction. A combination of good system design practices, good circuit layout, good equipment layout, and, if necessary, a small amount of circuit filtering, will usually suppress interference to acceptable levels and eliminate various types of interaction phenomena. The following system considerations are recommended:

1) Operate parallel and potentially interacting SCR circuits from a stiff (low-reactance) supply line

2) If supply line is soft (high-reactance), consider using separate transformers to feed the parallel SCR branch circuits. Each transformer should be rated at no more than the required rating of the branch-circuit load

3) Avoid purely resistive loads operating from stiff lines; they give highest rates of current rise on switching

4) Keep both leads of a power circuit wiring run together; avoid loops that encircle sensitive control circuitry

5) Arrange magnetic components to avoid interacting stray fields

Section II. FILTERS, CAPACITORS, AND INDUCTORS

3-6. Filters

a. Filters are combinations of circuit components designed to pass currents at certain frequencies and to attenuate currents at other frequencies. They utilize the resonance characteristics of series and parallel combinations of inductance and capacitance. These reactances reduce interference by introducing a high impedance in series with the interference currents and/or shunting interference currents to ground through a low impedance. Figure 3-52 shows the attenuation-frequency curves for four common types of filters and figure 3-53 shows typical filter arrangements.

b. The most common types of interference filters are the L-section, the T-section, and the π-section. The π-type filter is the most widely used because it provides greater attenuation than do the L- and T-types for only a small increase in size. All of the filters shown on figure 3-54 are low-pass filters. To convert them to high-pass filters, all of the L's are replaced by C's, and all of the C's are replaced by L's.

(1) The insertion loss, α, for the filters of figure 3-54 may be calculated from the following equations:

$$\text{Insertion loss} = \alpha = 20 \log_{10} \left| \frac{E_1}{E_2} \right| \qquad (3\text{-}34)$$

where: E_1 = load voltage without filter in the circuit

E_2 = load voltage with filter in the circuit

(2) For the L-type filter of figure 3-54A:

$$\text{Let } R_1 = R_L = R$$

$$\text{Insertion loss} = 10 \log_{10} \left[\frac{(2-\omega^2 LC)^2 + (\omega CR + \frac{\omega L}{R})^2}{4} \right] \qquad (3\text{-}35)$$

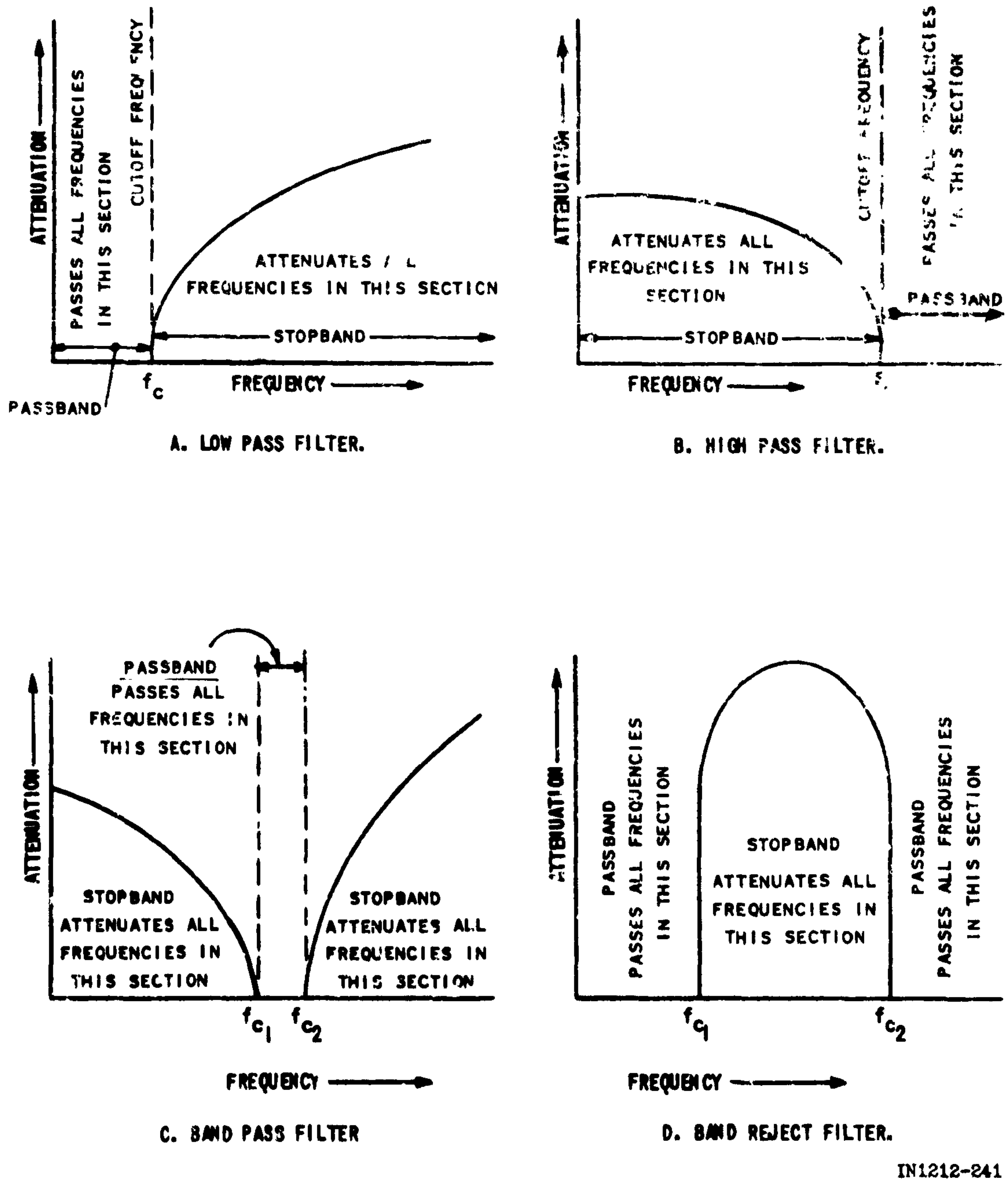
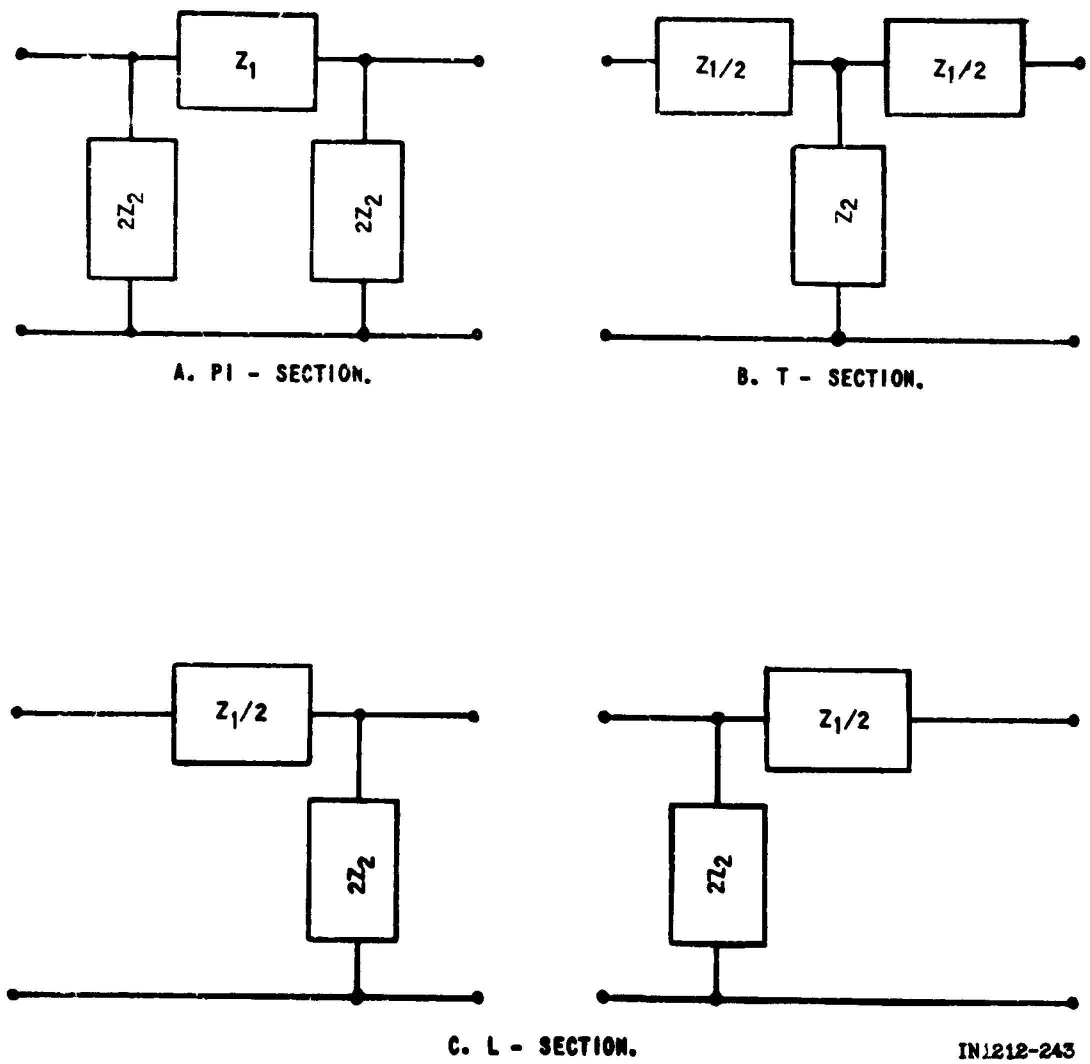
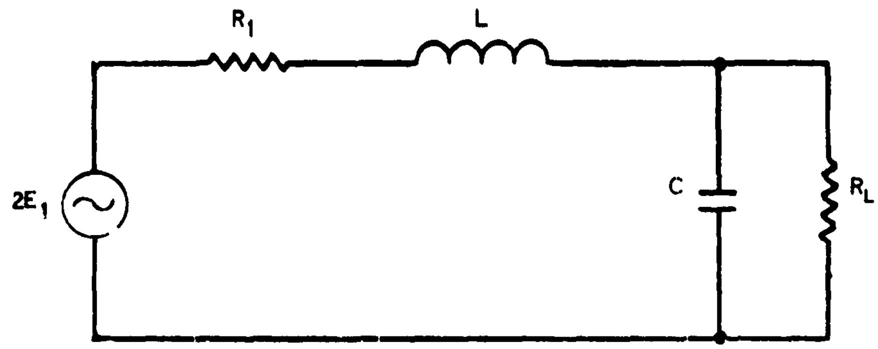


Figure 3-52. Filter Attenuation - Frequency

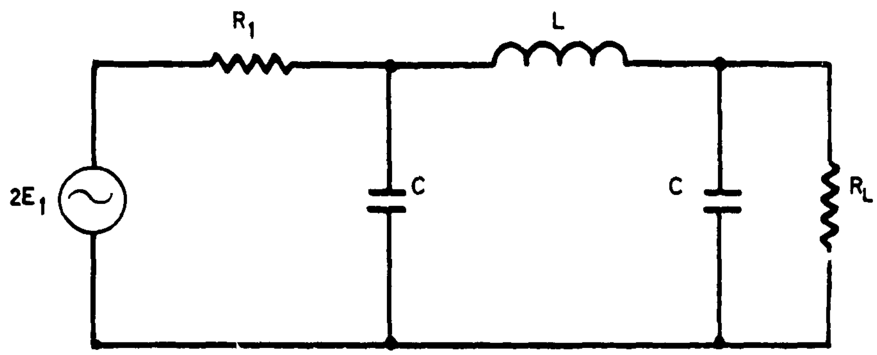


IN1212-243

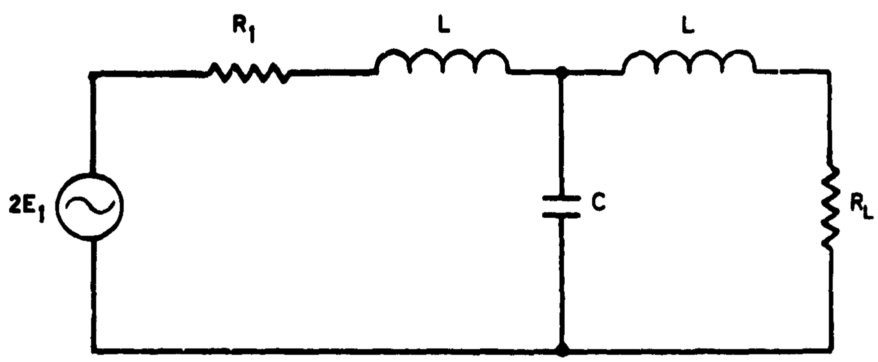
Figure 3-53. General Types of Filter Networks



A. L-TYPE FILTER.



B. π-TYPE FILTER.



C. T-TYPE FILTER.

IN1212-244

Figure 3-54. Representative Low Pass-Filters

(3) For the π-type filter of figure 3-54B:

$$\text{Let } R_1 = R_L = R$$

$$\text{Insertion loss} = 10 \log_{10}\left[(1-\omega^2LC)^2 + \left(\frac{\omega L}{2R} - \frac{\omega^3LC^2R}{2} + \omega CR\right)^2\right] \quad (3\text{-}36)$$

(4) For the T-type filter of figure 3-54C:

$$\text{Let } R_1 = R_L = R$$

$$\text{Insertion loss} = 10 \log_{10}\left[(1-\omega^2LC)^2 + \left(\frac{\omega L}{R} - \frac{\omega^3L^2C}{2R} + \frac{\omega CR}{2}\right)^2\right] \quad (3\text{-}37)$$

3-7. Filter Characteristics

a. Ratings. Filters are usually inserted in a circuit so that all circuit energy passes through them; they must therefore perform their functions without impairing normal circuit operation. Filters are generally rated in terms of the voltage and current parameters of the circuit in which they operate.

b. Attenuation and Insertion Loss.

(1) Attenuation and the frequency range of attenuation are the primary characteristics that determine filter suitability for interference reduction. If a selected filter does not provide the minimum attenuation required in the stop-band, it is not satisfactory, regardless of its other characteristics. Figure 3-55 illustrates a typical attenuation-versus-frequency curve for a power-line filter. The attenuation of the filter is expressed as the ratio of the filter input voltage to the filter output voltage, measured under normal circuit conditions:

$$\text{Attenuation (db)} = 20 \log_{10}\left|\frac{E_1}{E_2}\right| \quad (3\text{-}38)$$

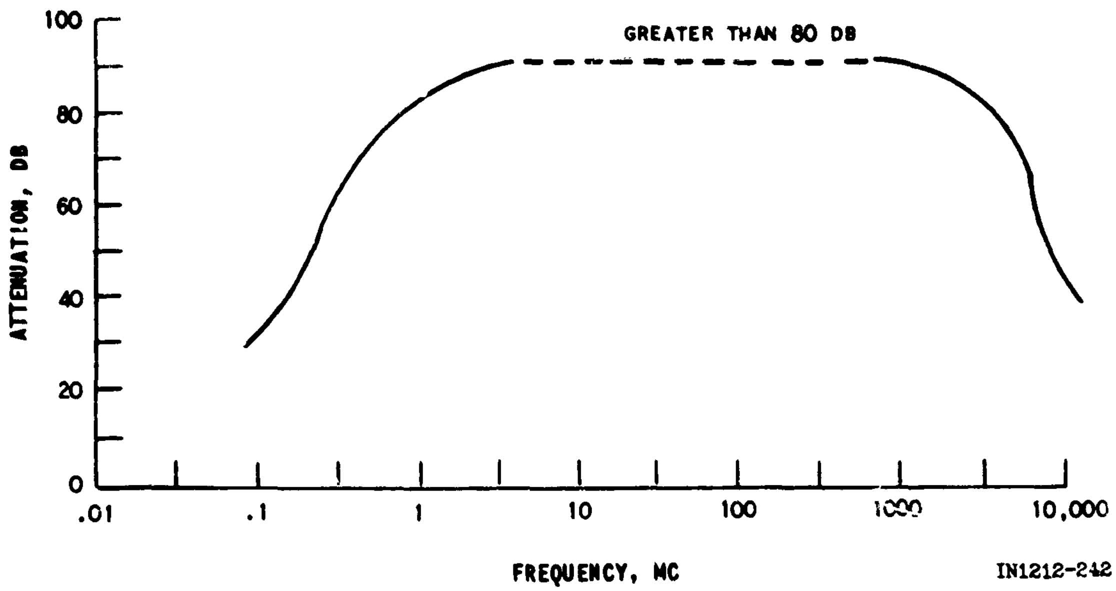


Figure 3-55. Attenuation vs Frequency for a Power Line Filter

where: E_1 = voltage across filter input terminals

E_2 = voltage across filter output terminals

(2) The attenuation figure, however, does not take into consideration the source and load impedances and, therefore, does not represent a true indication of the suppression effectiveness of a filter. The use of the insertion loss criteria is a far more realistic measure of a filter's effectiveness, as it is a function of both source and load impedances as well as a function of the filter network itself.

(3) In their filter catalogs, most filter manufacturers quote values of insertion loss (the ratio of voltages at a given frequency appearing across the load terminals before and after the filter is inserted into a circuit):

$$\text{Insertion loss (db)} = \alpha = 20 \log_{10} \left| \frac{E_3}{E_2} \right| \qquad (3\text{-}39)$$

where: E_2 = load voltage with the filter in the circuit

E_3 = load voltage without the filter in the circuit

Insertion loss is usually quoted by manufacturers for a 50-ohm system. If the circuit to be filtered does not have both a 50-ohm input and output impedance, the insertion loss will differ from the catalog value. In these cases it is advisable to use matching networks to make the system 50 ohms.

c. Filter Considerations. The following characteristics are common to all filter installations and should be carefully considered in filter selection:

1) Voltage rating of the circuit in which the filter is to be inserted

2) Maximum current that will pass through the filter

3) Duty cycle of the filter; this applies to the decreased load current of intermittent operation

4) Operating frequency of the circuit and the frequencies to be filtered

5) Voltage drop that can be tolerated at the operating frequency

6) Maximum ambient temperature at which the filter must operate

7) Attenuation required of the filter for adequate interference reduction

8) Minimum filter life -- the number of hours a unit will operate satisfactorily under rated conditions and at maximum ambient temperature

9) Circuit requirements such as minimum (or maximum) capacitance or insulation resistance:

a) Voltage and current ratings required of a filter, unless otherwise specified, are the maximum allowable for continuous operation. Any filter will perform satisfactorily when operated below its nameplate rating. The breakdown voltage of capacitors used in filters should also be considered. A safety factor of approximately 100-percent should be used. For a given application, the working voltage of a standard filter capacitor should be twice the voltage of the circuit in which it is used. In general, filter test voltage should be twice the filter's nameplate rating

b) Power-frequency specifications are primarily applicable to low-pass line filters. Filters should not be operated at power frequencies above those specified by the manufacturer. They will operate satisfactorily at frequencies below those marked on the nameplates

c) All filters using series inductors cause some voltage drop. The magnitude of this drop is determined by the series resistance of the filter and may be expressed, for example, as: maximum voltage drop at rated current = 0.1 volts. In some cases, only the series resistance is given. When this occurs, the voltage drop for dc filters, or ac filters when the cut-off frequency is far removed from the operating frequency, can be calculated using Ohm's law

d) If the ambient temperature deviates from the range specified for the filter, failure or shortened service life may result

e) The capacitance to ground of a filter is often a determining factor in the filter's application. Some circuits may limit the maximum capacitance to ground to prevent long time-constant currents from charging the capacitors through a resistor. Another instance in which capacitive limitations are important is when capacitors might cause danger to personnel because of charging currents

f) The insulation resistance of a filter decreases continually during the life of the filter. The resistance of a new filter usually measures several hundred megohms, and, after several years of operation, may measure 50 megohms. Because most power cables can be used with an insulation resistance of one megohm, the insulation resistance of a filter has little effect on such cables

g) The following nameplate information, taken from an actual filter, is typical of most filter ratings:

RF Interference Filter SP-99

For Telephone Line Application

Current: 2 x 0.15 Amperes

Voltage: 250 VAC

AC Frequency: 60 CPS

Voltage Drop At Maximum Frequency: 0.1 volts
(at unity P.F. for AC units)

Duty Cycle: Continuous

Operating Tempera..e Range: -55°C to +85°C

Weight: 2 pounds, 8 ounces

Test Voltage: 1000 VDC

Special
Features: Nominal 600 ohms pass-band impedance

Less than 0.5 db pass-band insertion loss
from 50 cps to 15 kcs in 600-ohm line

60 db at 50 kcs, measured per MIL-STD-200

Hermetically sealed

Hot-tinned nonferrous case

Pressurized mounting

Glass or ceramic terminals

d. Multiple Circuit Filter Versus Single Units. The problem of equipment requirement for many filters has two possible solutions. The first is the use of a single, hermetically sealed box containing all the LC circuit components needed for filtering the equipment. Input to the container may be accomplished by compression solder-sealed terminals; output, by a connector through a bulkhead. The advantages of this method of packaging are:

1) Field replacement is comparatively easy, and the down-time required to return the inoperative equipment to use is short

2) All power leads are routed to one point, passed through a black box, and out of the equipment through a single point

3) A small saving in size and weight is effected

Its disadvantages are:

1) Failure of any circuit means failure of the entire unit, since replacement of component parts within the hermetic enclosure is not practical

2) The reliability of this type of unit is greatly reduced, compared to a single network, by the large number of components involved

The alternative solution is the use of single-circuit filters. The advantages of using single-circuit filters are:

1) Availability

2) Failure of one component does not mean the loss of all filters in the system

3) Reliability is increased by the reduction of the number of components per enclosure

4) Cost and delivery time may be reduced

The disadvantages of this solution are:

1) More complex harnesses may be needed

2) Size and weight may be increased

A combination of the two methods, if space limitations permit, offers the most practical solution. Individual standard components can be located within an open framework to provide replaceable units within a common enclosure. The output leads can be brought to a connector located on the frame to provide the single exit point.

3-8. Lossy Transmission Line Filter

a. The usual π or T filters, intended for broad-band interference filtering, are generally composed of lossless or very nearly lossless, inductive and capacitive lumped elements. Such filters cannot dissipate energy within their rejection range; they merely reflect it, reroute it, or transform it so that, under certain conditions, it may reappear elsewhere as an undesirable signal or interference. For any configuration of lossless circuit elements of a filter, there may be found, for any frequency, a load impedance for which the filter will transmit to the load the maximum energy available from the source. Given the proper source and load impedances, insertion of a lossless filter may actually increase the energy delivered to the load; or the filter, in that circuit at that frequency, might have negative insertion loss. Similarly, the application of a filter composed of lossless lumped elements, may also, under proper circuit impedance conditions, increase the voltage or current at the load. Unlike carefully designed laboratory circuits used for insertion loss measurements, where source and load impedances are fixed at exactly 50 ohms resistive, the impedances that a filter sees in most practical power line applications are extremely variable with frequency, and range from very high or very low resistive to nearly minus or plus infinity reactive. In a practical application, it is therefore not unlikely that at one or several frequencies within the range over which the filter is expected to be effective, the circuit impedances will cause a critical lowering of the filter's insertion loss. These impedances may even render the insertion loss altogether negative. Demonstrations of this effect have been observed in which application of a reactive filter to a line carrying interference has actually resulted in more, rather than less, interference voltage appearing on the line beyond the point of application. This deficiency, inherent in all filters composed of lossless elements, has led to the investigation of a dissipative type of filter that takes advantage of the loss versus frequency characteristic of dielectric materials such as ferrites.

b. The dissipative filter is a short length of ferrite tube with conducting silver coatings deposited in intimate contact on the inner and outer surfaces to form the conductors of a coaxial transmission line. The line becomes extremely lossy; that is, it has high attenuation-per-unit-length in the frequency range where either electric or magnetic losses, or both, become large and increase rapidly with frequency. Figure 3-56 illustrates a typical insertion loss versus frequency curve for a lossy ferrite tube transmission line filter. Dissipative filters of this type are necessarily low pass; and the large field of application is general purpose power line filtering.

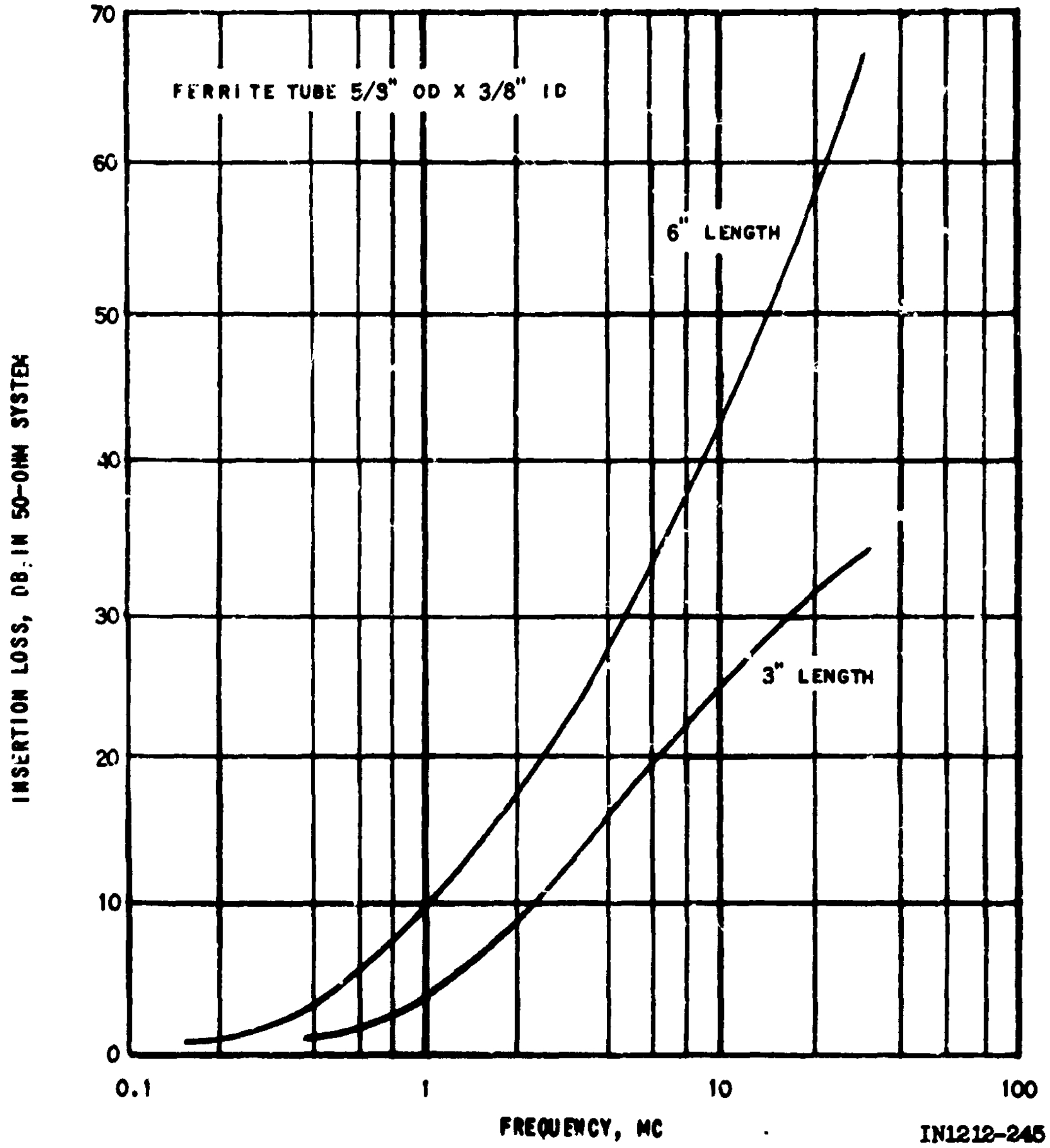


Figure 3-56. Lossy Ferrite Tube Filter Insertion Characteristics

3-9. Microwave Filters and Matching Networks

a. Requirements for microwave filters are often met by coaxial or waveguide construction. The coaxial technique is applicable to 2 or 3 kmc, while waveguide elements are normally used above these frequencies. The upper frequency limit for coaxial line filters is usually determined by several factors, among which are fabrication tolerances and construction arrangement. Resonant waveguide elements are useful for narrowband filters, but wideband filter requirements must usually be met with cascaded waveguide elements. Beginning in the early 1950's, stripline techniques have been applied to the problem of fabricating microwave filters.

b. Band pass, band rejection, high pass, and low pass filters have been produced using strip transmission line. Figure 3-57 shows an example of one of several configurations of stripline presently in use. A single copper strip forms the center conductor. This strip is embedded between two dielectric sheets which are surrounded by two copper plates that form the ground plane. In practice, it has been found that filter construction using stripline limits the designer to simple but adequate combinations of series lines, open or shorted shunt lines, and series capacitors (formed by transverse slots). Shunt lines are formed by lengths of stripline at right angles to the mainline. Short lengths of open circuited line appear capacitive; short lengths of short circuited line appear inductive. A method of producing shorted shunt lines is shown in Figure 3-58. A common problem encountered in the use of stripline techniques is the possibility of propagating higher order modes. Such modes can be excited by any unintentional tilt of the center conductor. The result can be narrow spurious pass responses in a rejection band. These can be eliminated by loading the line with resistor cards, powdered iron slugs, or screws located so as to absorb energy from the higher modes without affecting the main transverse electromagnetic (TEM) lines. In addition to preventing the propagation of undesired higher-order modes by electrical means, the above methods provide a mechanically rigid structure.

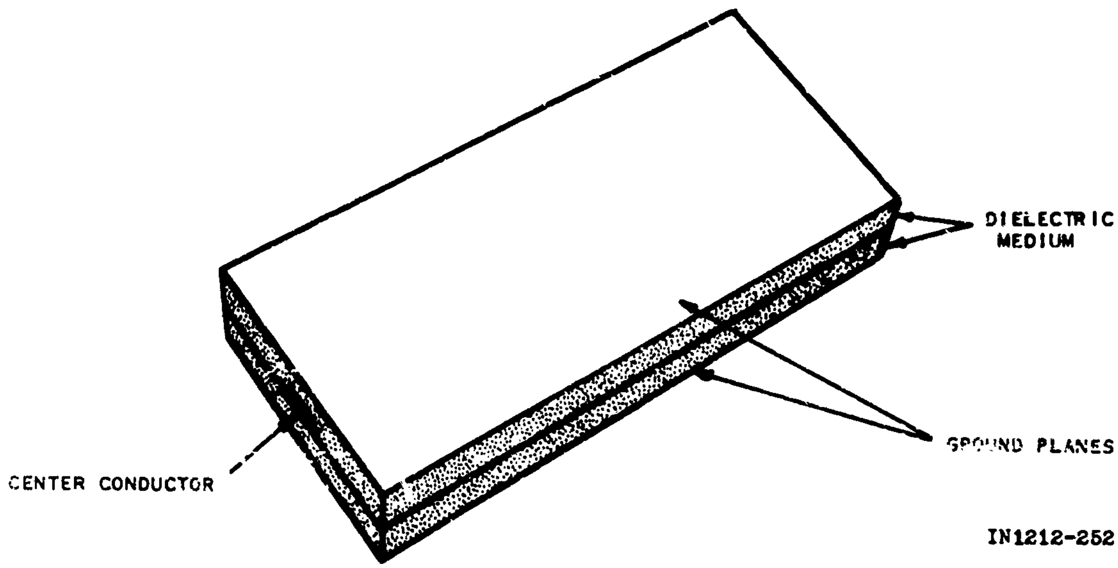


Figure 3-57. Strip Transmission Line Construction

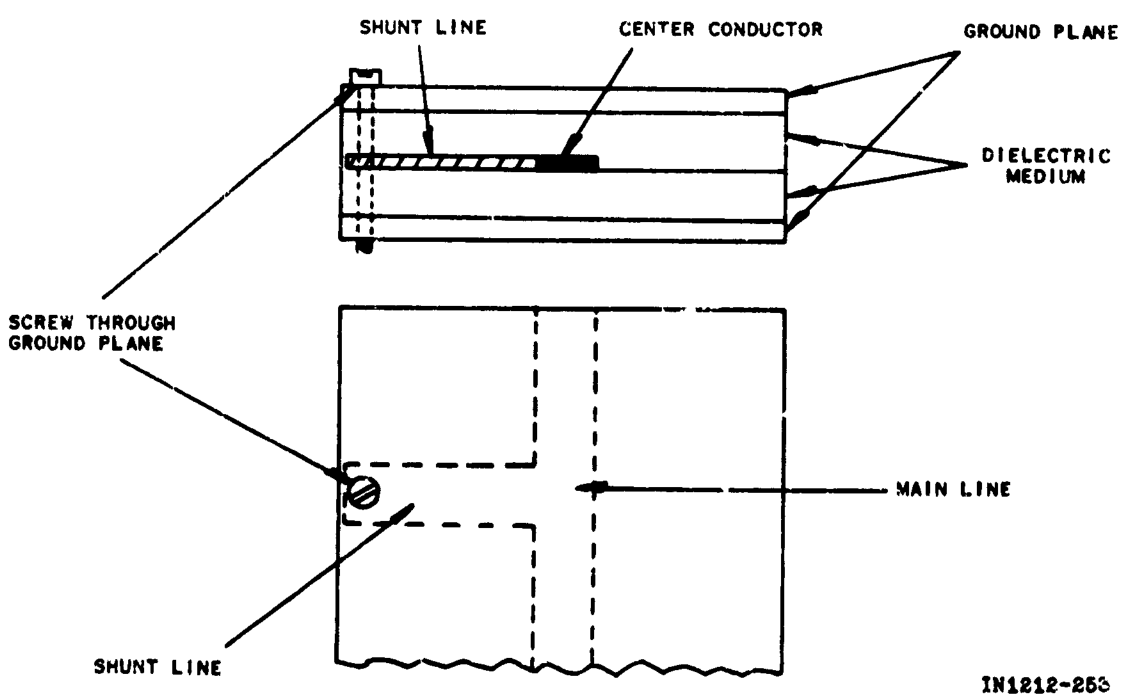


Figure 3-58. Shorted Shunt Line Construction

c. The commonly used π and T filter sections can be fabricated using stripline techniques. Figure 3-59 shows the stripline center conductor pattern and the low frequency equivalent circuit for a basic T-section. The inductances shown in the lumped circuit are represented in stripline form by short lengths of line terminated by transverse elements. The value of the inductive elements may be varied by altering the width, (W), of the center conductor and hence the characteristic impedance. The capacitive element in the lumped model is produced by a transverse line of the appropriate electrical length. Note that the transverse element is placed symmetrically about the center conductor instead of being placed totally on one side. Splitting the shunt capacitive reactance into two equal sections on either side of the line shortens the necessary physical length of the stubs and therefore places spurious responses farther away from cutoff and into the stop band. By combining the basic π and T sections, more complex filters may be fabricated. The center conductor configuration for a low pass filter and its equivalent low frequency network are shown in figure 3-60. Filters of this type are useful because they are comparatively small and spurious responses are at frequencies far removed from the pass band.

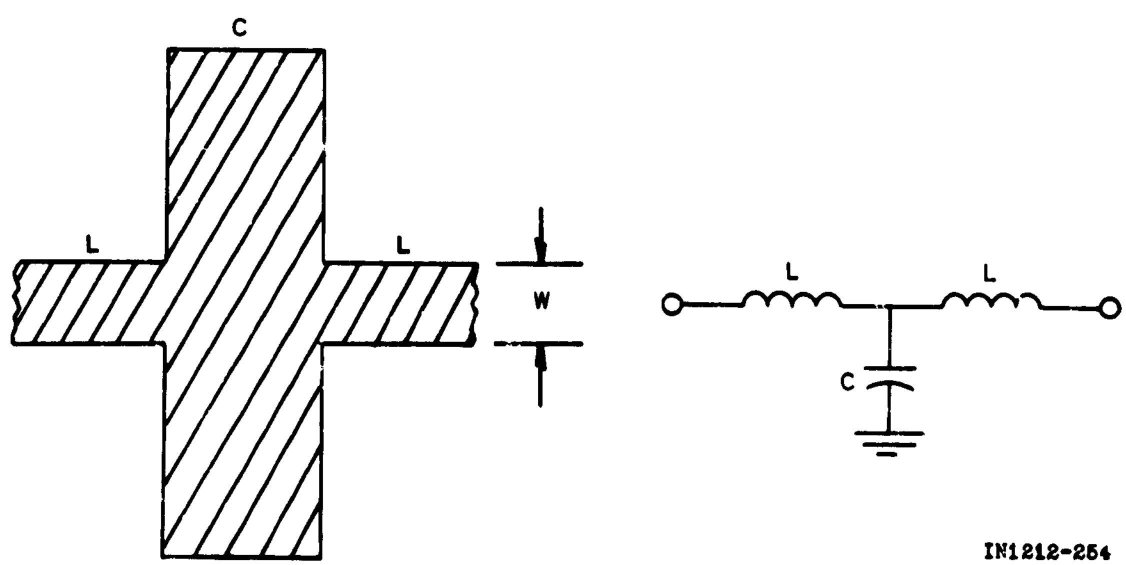


Figure 3-59. Symmetrical T-Section Filter and Equivalent Network

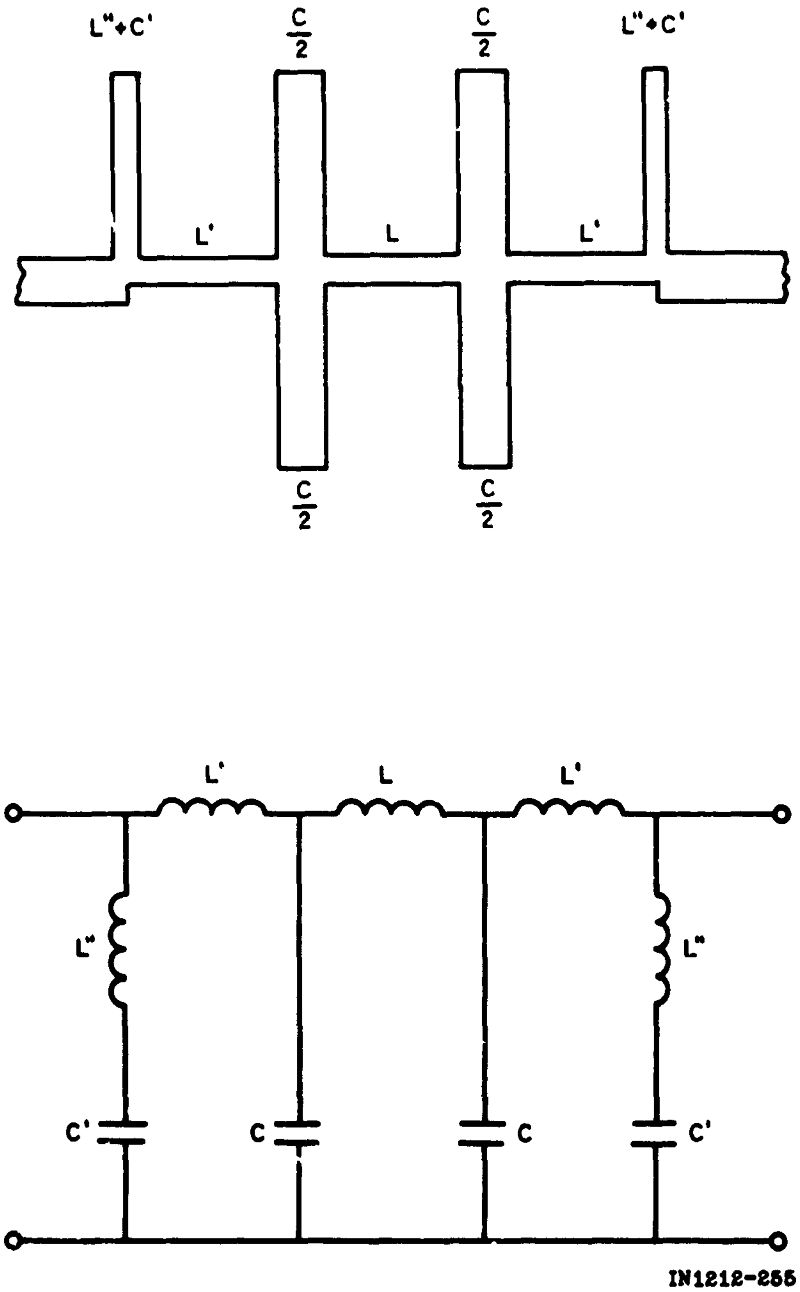


Figure 3-60. Low-Pass Filter and Equivalent Network

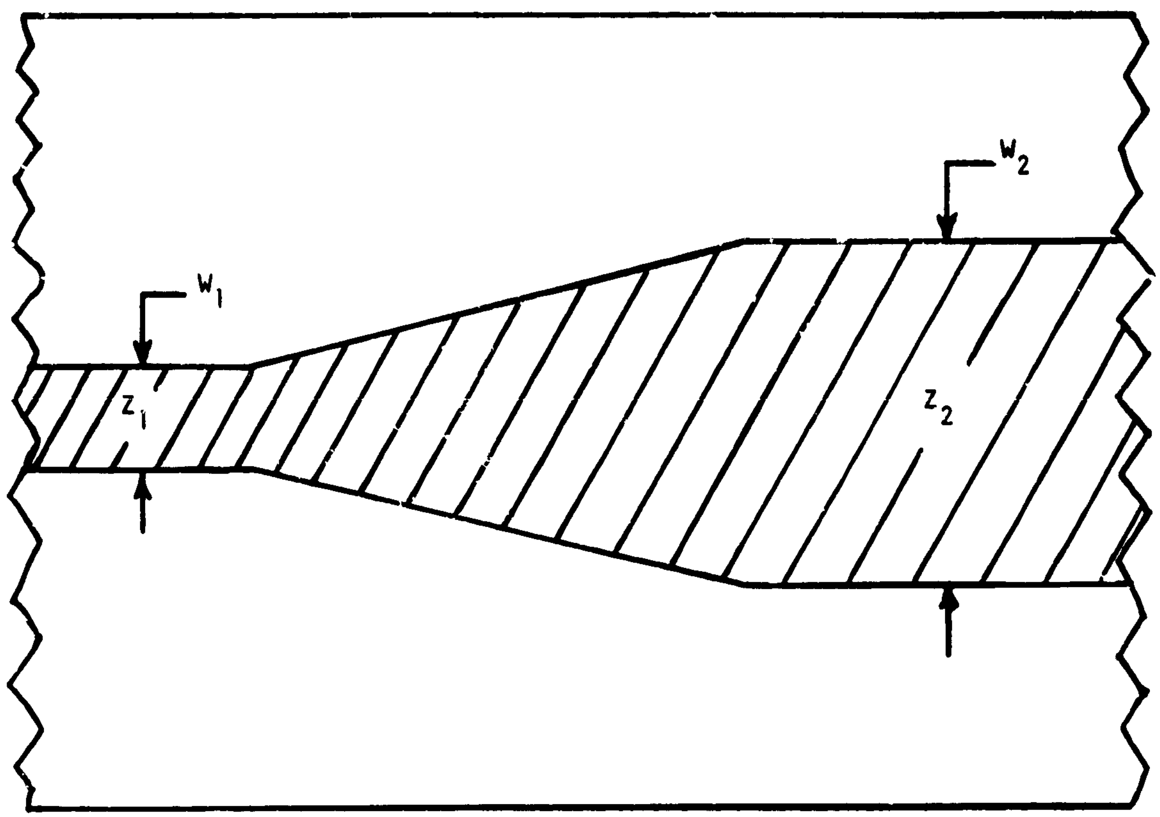
d. There are two types of spurious responses generally encountered in this type of stripline filter. One type of response occurs at a frequency where the spacing between shunt elements is equal to a half wavelength and at integer multiples thereafter. Another type occurs when the shunt element length equals a half wavelength or an integer multiple thereof. In carefully designed stripline filters, these responses occur far above the pass band. If it is found necessary, these responses may be eliminated by cascading the filter with a low-pass filter having a cutoff frequency slightly lower than the first spurious response and having no coincident spurious responses.

e. Impedance matching networks may be readily fabricated in stripline. Purely resistive matching can be accomplished by linearly or exponentially tapering the width of the center conductor. Figure 3-61 shows a line linearly tapered to match two different characteristic impedances, Z_1 and Z_2. More complex matching may be accomplished by including shunt reactive elements of the proper electrical length.

f. The relative simplicity of fabricating stripline circuits and the inherent low losses make the use of stripline attractive for microwave applications.

3-10. Waveguide Filters

All waveguides act as high-pass filters with cutoff frequencies (the lowest frequencies at which propagation will occur without attenuation) determined by the shape and size of the waveguide, and by the mode of transmission. The greater the number of joints and bends in



IN1212-309

Figure 3-61. Linear Taper Matching of Strip Lines

a system, the greater its capacity for creating and supporting different transmission modes. The attenuation characteristics of hollow-pipe guides are similar to those of conventional high-pass filters in which the electrical elements are lumped rather than distributed. At frequencies below the cutoff frequency, power is rejected because the guide dimensions and frequency do not permit the existence of the type of waves that transmit power. In this region, there is no real propagation, and the fields are attenuated exponentially. The attenuation (α) in a length of guide (L) is:

$$\alpha = 54.5 \frac{L f_c}{v_c} \left[1 - \left(\frac{f}{f_c}\right)^2 \right]^{\frac{1}{2}} \text{ decibels} \qquad (3\text{-}40)$$

where: f_c = cutoff frequency

f = operating frequency

v_c = velocity of light (v_c must have the identical length units as L)

In the design of equipment enclosures, it is almost always necessary to provide openings for control shafts, meters, and ventilation. Such openings act as windows for radiated interference. An extremely useful interference reduction technique is to design the aperture through which leakage occurs to be a waveguide-type attenuator. Values of attenuation for circular and rectangular waveguides are given on figures 3-62 and 3-63.

a. Wide-Band Reflective Waveguide Filters. The serrated-ridge type waveguides provide high stop-band insertion loss over a wide frequency range. Values of pass-band losses of less than 0.1 db, and of stop-band losses greater than 50 db over a two-to-one frequency range, are easily attained with these waveguides. The power-handling capacity of this type of waveguide filter is only about one per-cent of the unperturbed waveguide. If carefully evacuated, such filters are able to operate at up to half the power-handling capacity of corresponding unridged waveguides at one atmosphere of air pressure. This type of filter may also be useful as part of a hybrid filter where the power is divided into several branches. A similar corrugated waveguide filter (fig. 3-29 in Section 1 of this chapter) has been designated the waffle-iron filter because of its construction. As in serrated, ridged waveguides, only the dominant mode propagates in the stop-band; therefore, there is no multimode problem. This filter is very compact, easily fabricated, and has a power handling capacity of about three per-cent of the corresponding uncorrugated

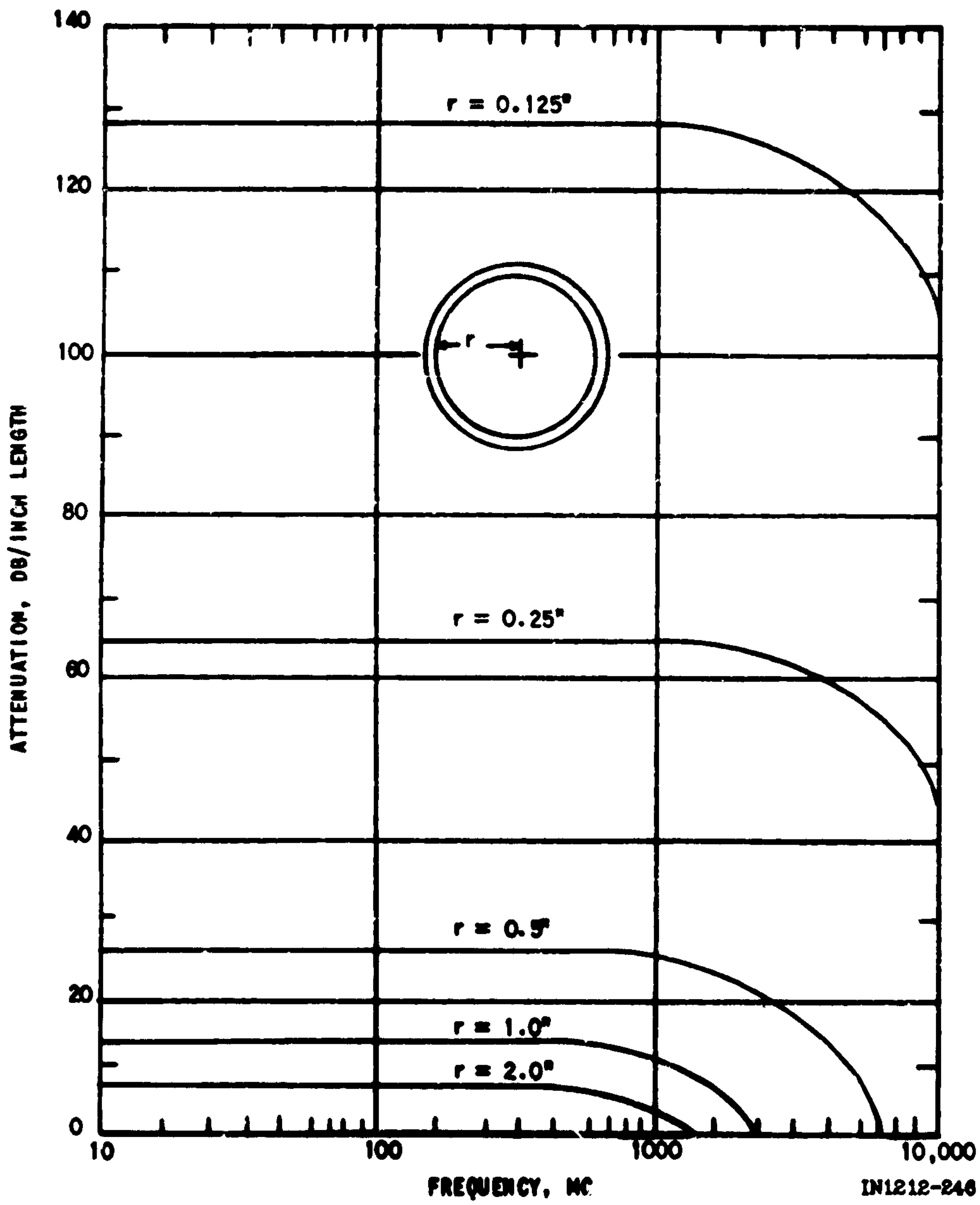


Figure 3-62. Attenuation of a Circular Waveguide for TE_{11} Mode

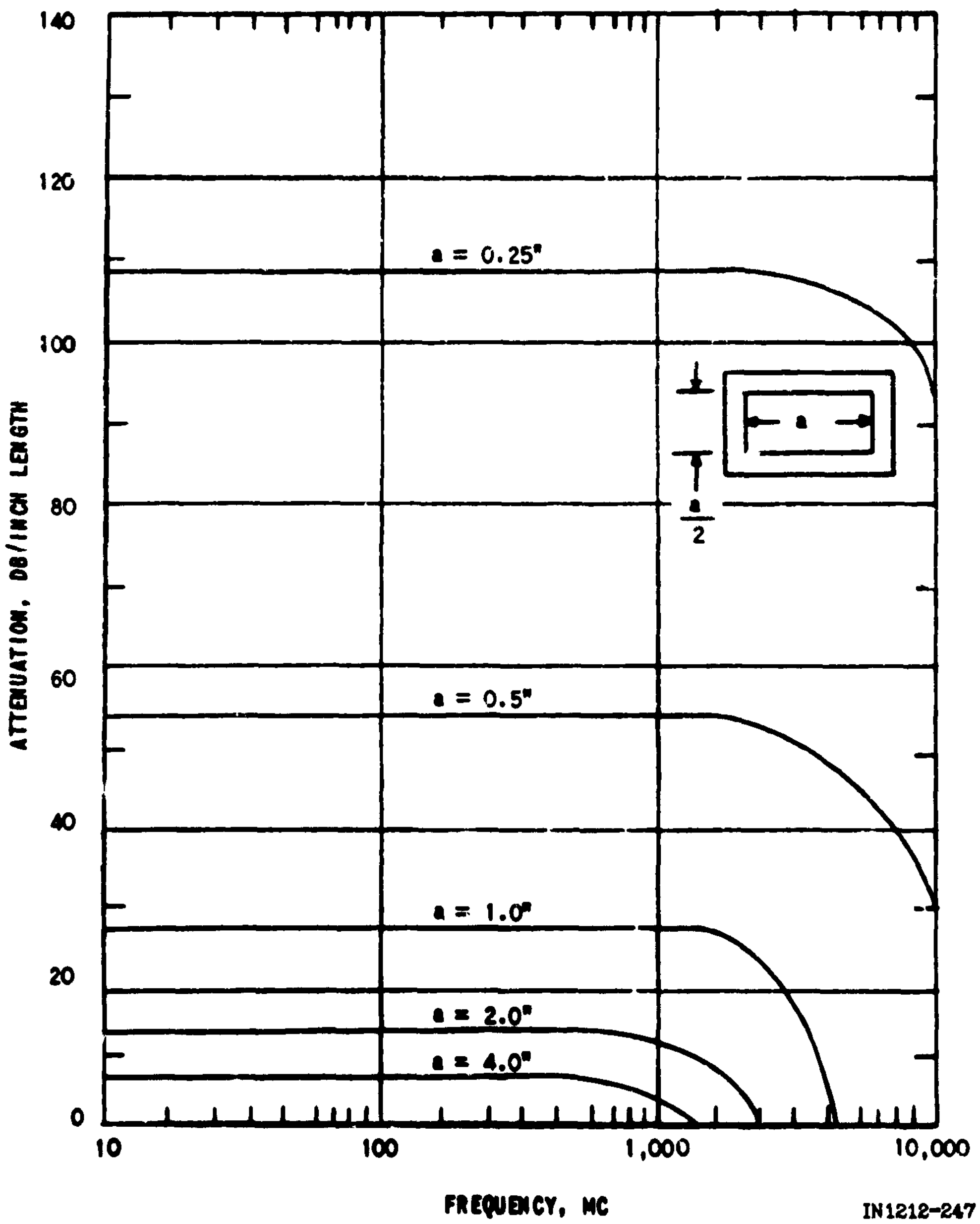


Figure 3-63. Attenuation of a Rectangular Waveguide for TE_{10} Mode

waveguide. The same filter, when evacuated, is capable of handling 100 per-cent of the power that the unperturbed waveguide can handle at one atmosphere of air pressure. Figure 3-30 in Section 1 of this chapter shows the characteristics of this filter.

b. Reactive Mode Devices. There are numerous reactive devices that can be used in filter designs to increase the stop-band insertion loss for a particular mode. Decreasing the height dimension of the waveguide in the filter will reflect most of the TE_{0n} modes and the degenerate modes, thereby providing high insertion loss (50 db or more). However, the power handling capacity is decreased in the same proportion that the height dimension is decreased. The same insertion loss for the same modes can be attained by placing a septum in the waveguide, parallel to the Y-Z plane, which does not disturb the propagation of the dominant mode. By tapering and rounding the edges, as shown on figure 3-64, the gradient field enhancement at the front and back edge of the septum becomes negligible. Thus, a thin septum has little effect on the power handling capacity. Septum sections have been high-power tested and shown not to degrade the system. Septa can be constructed either in the form of a series of rods projecting across the filter, or in the form of bullets projecting out from the narrow walls (fig. 3-65).

c. Tuned Cavities. Another reactive device that is useful for increasing the stop-band loss is the tuned cavity. Figure 3-66 shows a section of such a cavity attached to a waveguide section. A few cavities can add 30-db loss over a narrow frequency range. Placement of these cavities away from the regions of maximum electric field enables the section to maintain 100 per-cent power handling capacity.

d. Ferrite Filters. The use of resonant ferrite materials in waveguides to provide the desired attenuation at certain frequencies has been discussed and proposed by a number of investigators. Ferrite materials, properly located in a waveguide, provide attenuation by absorbing undesired energy, thus providing a nonreflective filter. By placing a thin slab of ferrite material across the broad walls of a section of waveguide and biasing the ferrite with magnetic fields that cause resonances to occur, it is possible to absorb large amounts of microwave power over a selected frequency range. In using ferrites to absorb harmonic signals, the pass-band to stop-band loss maintains a ratio similar to, but higher than, that of ferrite isolators. A substantial fundamental power loss may be experienced in attempting to obtain extra broad stop-band loss. Ferrite slabs, placed on the waveguide broad walls, will not absorb energy from all modes with equal efficiency as the intensities of the rf magnetic fields often vary considerably from mode to mode. In such filters, it is often necessary to reduce the height dimension of the waveguide to generate a sufficient magnetic field across the ferrite slab. In so doing, a sizable amount of the energy in the narrow wall modes is reflected instead of absorbed, and the power-handling capacity is lowered. Excellent characteristics of 0.2 db to 0.3 db pass-band loss and 30 db to 50 db stop-band loss can be achieved using ferrites at powers in the range of five megw and higher.

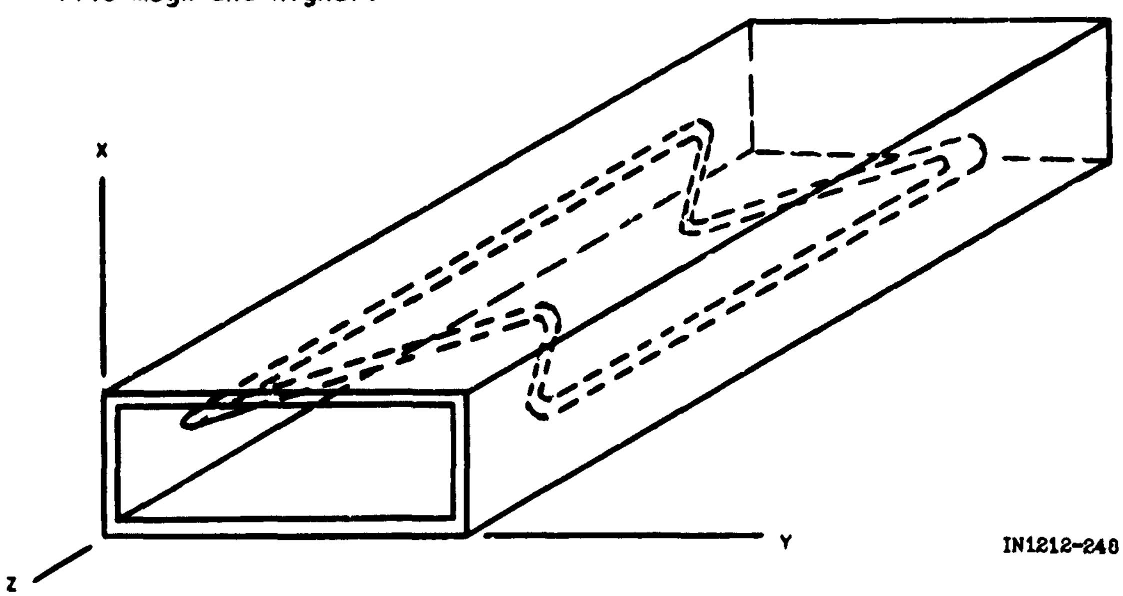


Figure 3-64. Septum Section

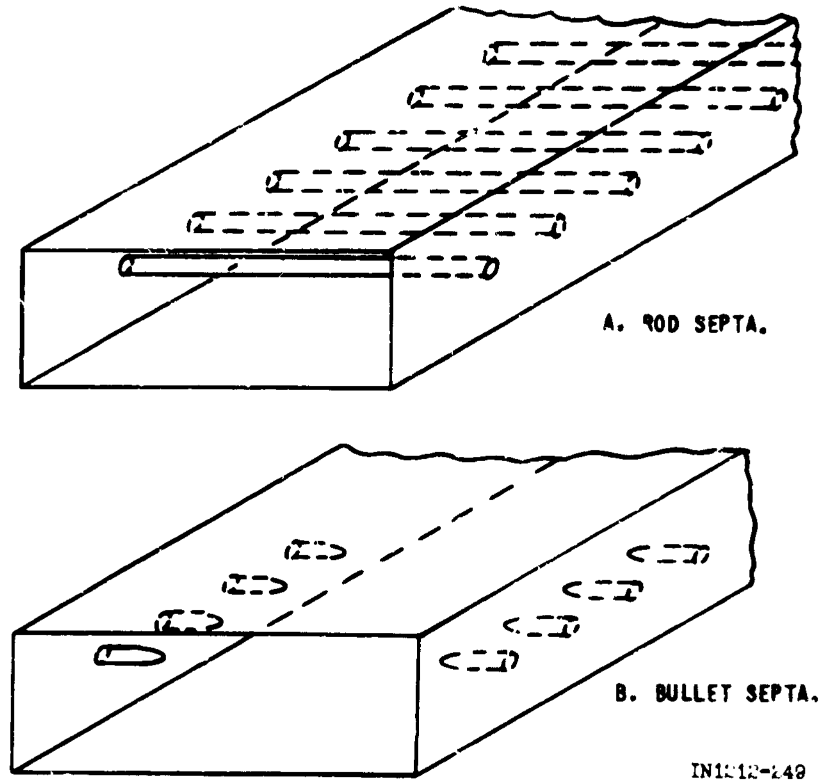


Figure 3-65. Rod and Bullet Septa

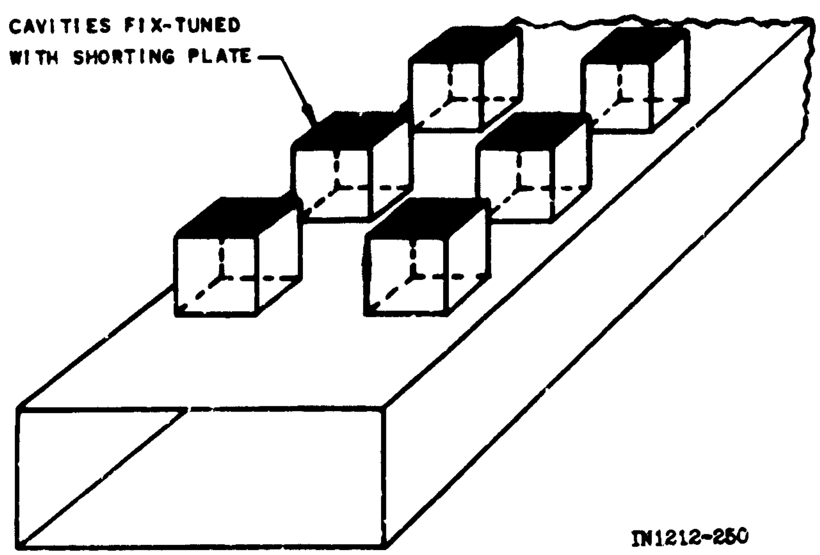


Figure 3-66. Tuned Cavity Section

e. Mode Filters. There are various filter techniques that can be used to absorb certain modes. Among these techniques is the placement of a thin resistance film perpendicular to the narrow wall and parallel to the broad wall of a waveguide, as shown on figure 3-67. Such a film effectively absorbs many of the modes that have a narrow wall component of electric field, without affecting the dominant mode. A narrow slot, filled with absorbing material and placed along the broad wall of the waveguide in a manner similar to a slotted line, will absorb the energy from all modes which have a current path perpendicular to the slot.

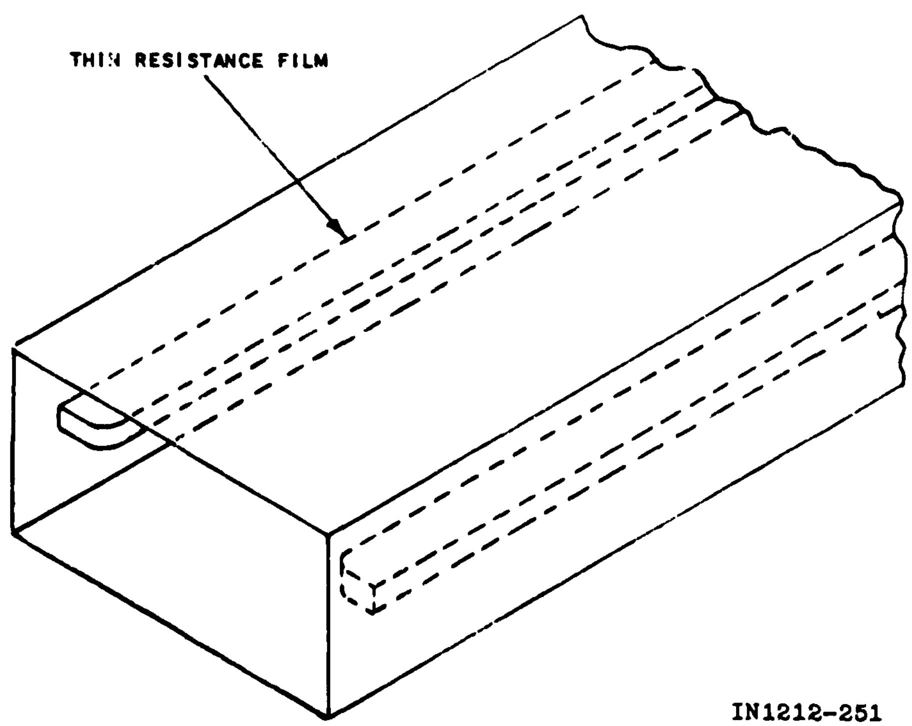


Figure 3-67. Narrow Wall Mode Absorber Strips

3-11. Filter Installation and Mounting Techniques

a. General.

(1) Electromagnetic interference can emanate from equipment both by radiation through space and by conduction through power lines and control circuits. No matter how well-shielded a source may be, the shielding effectiveness can be nullified by conduction of interfering currents in the power, control, and instrumentation leads. Capacitors and filters prevent interference from reaching other circuits by introducing high-impedance paths for the interfering currents, or shunting them from the load through a lower impedance to ground, or providing a reflective mismatch to the interference.

(2) When using filters, proper installation is absolutely necessary to achieve good results. Effective separation of input and output wiring is mandatory, particularly for good high-frequency performance, because the radiation from wires carrying interference signals can couple directly to output wiring, thus circumventing and nullifying the effects of shielding and filtering. Input and output terminal isolation is most easily accomplished by using a filter that mounts through a bulkhead or chassis. In all cases where bulkhead mounting isolation is not feasible, isolation by shielded wiring is mandatory. It is highly desirable to locate suppression components in or on the device generating the interference. The rf impedance between filter-case and ground must be as low as possible. The methods of mounting a filter become very critical at high frequencies. If complete isolation is effected between input and output, filter insertion loss will approach the design figure.

(3) The impedance to ground of a filter that is improperly mounted, from the standpoint of rf grounding, can become sufficiently large in value to cause interference voltages to develop across the impedance at radio frequencies. These voltages reduce the effectiveness of the filter. An important factor in filter performance is the bonding of the filter-case to the ground plane structure of the interference source. This requirement is of utmost importance if the filter is to achieve its design performance capability. Figure 3-68 illustrates the effect of poor bonding on a π-section filter. The path of rf currents is indicated by the arrows. When a poor bond exists, the filter-case is raised above ground by an impedance, Zb. With this condition existing, the current through the first shunt capacitor, C_1, divides at the junction of the two capacitors and Zb. Some curr[illegible] flows through Zb, depending upon the impedance. The remaining current flows through the second shunt capacitor, C_2, to the load, thus compromising filter performance. It is imperative that the surface on which a filter is mounted, as well as the mounting surface of the filter itself, be clean and unpainted. The rf impedance of the filter-can to ground should be as nearly zero as possible. If the surfaces are aluminum, and a good bond is required, the surfaces should be iridited, never anodized. The mounting ears, or studs, must ensure firm and positive contact over the entire area of the mounting surface. Although the location of the filter depends on the individual application, in general, it should be installed as close as possible to the interference source. In cases where there is no control over the interference source, the filter should be installed at the point of susceptibility. Grounding plays an important role in the application of filters and capacitors, whether the two-terminal by-pass

type or the three-terminal feed-through type is used. Grounding is an important factor in the case of feed-through capacitors and filters because these devices are inherently more effective in the high-frequency ranges than two-terminal capacitors, and there is, therefore, more to lose by excessive impedance in the ground circuit. When using by-pass capacitors for interference reduction purposes, only metal-cased by-pass capacitors should be employed. The cases should always be grounded directly to the chassis, either by suitable clamps or by threaded-neck type construction. Grounding by-pass capacitors by pigtail leads simply adds additional series inductance, thus lowering the resonant frequency and the usefulness of the by-pass capacitor.

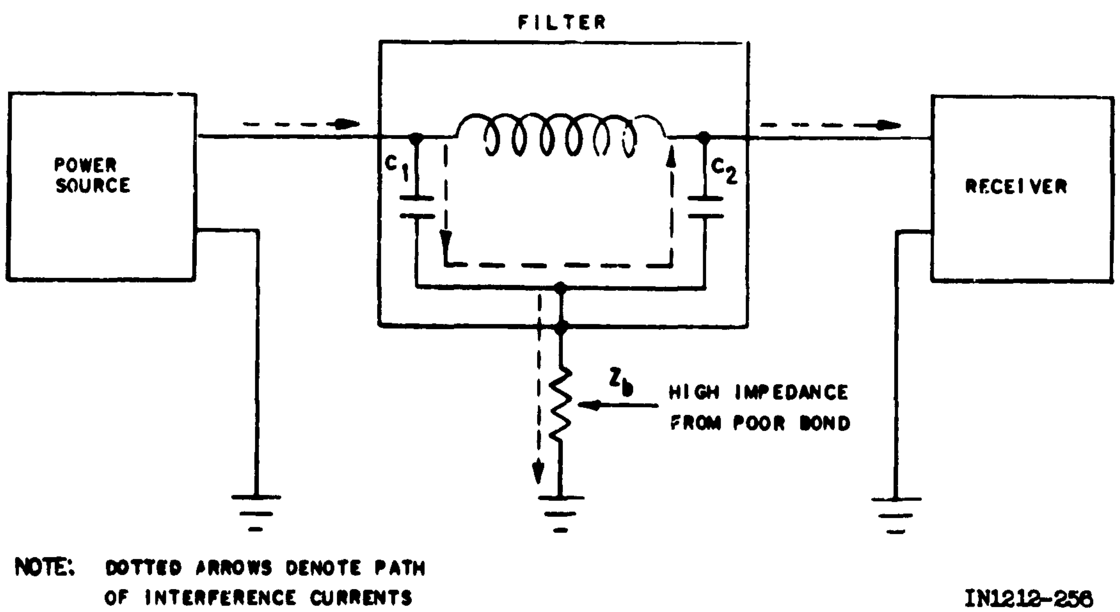


Figure 3-68. Effect of a Poor RF Ground Bond on Filter Effectiveness

b. Chassis Mounting. In general, any of the following five methods can be employed to mount filters on a chassis:

1) Tabs on the filter body
2) Screws or bolts on the filter body
3) A flange on the filter body
4) A clamp on the filter body
5) A feed-through stud for bulkhead mounting

Figure 3-69 is representative of filter mounting techniques. A typical filter installation is shown on figure 3-70. The top figure (3-70A) illustrates the preferred method of filter installation; one that is integral with the interference source -- in this case, a dc motor. Figure 3-70 also illustrates the difference between proper and improper filter installation. In 3-70A, the bulkhead mounting principle is used, and the filter's input and output circuits are completely isolated. Figure 3-70B shows how direct input-output coupling can reduce the effectiver of the filter, particularly when extremely high magnetic fields exist within the shielded area. This results in interference being coupled to all leads entering or leaving the area. Often, filters fail to perform because of improper installation. Poor installation can result from improper lead routing (fig. 3-71). In this figure, two incorrect methods of mounting a filter, both ineffective at high frequencies, are illustrat In figure 3-71A, the input and output leads are physically crossed, completely nullifying the effectiveness of the filter. In figure 3-71B, isolation between input and output circuits is not complete due to lack of shielding on the leads -- although there is the advantage of ease of assembly and some isolation up to about 5 mc. The insertion loss drops rapidly because of the coupling of energy across the filter, regardless of the insertion loss originally designed into the filter. Proper installation of power line filters is shown on figure 3-72B. One method of achieving designed insertion loss is to shield either the input or output lead, or both. To be really effective, shielded wire must be continuous from the interference source to the filter and/or from the

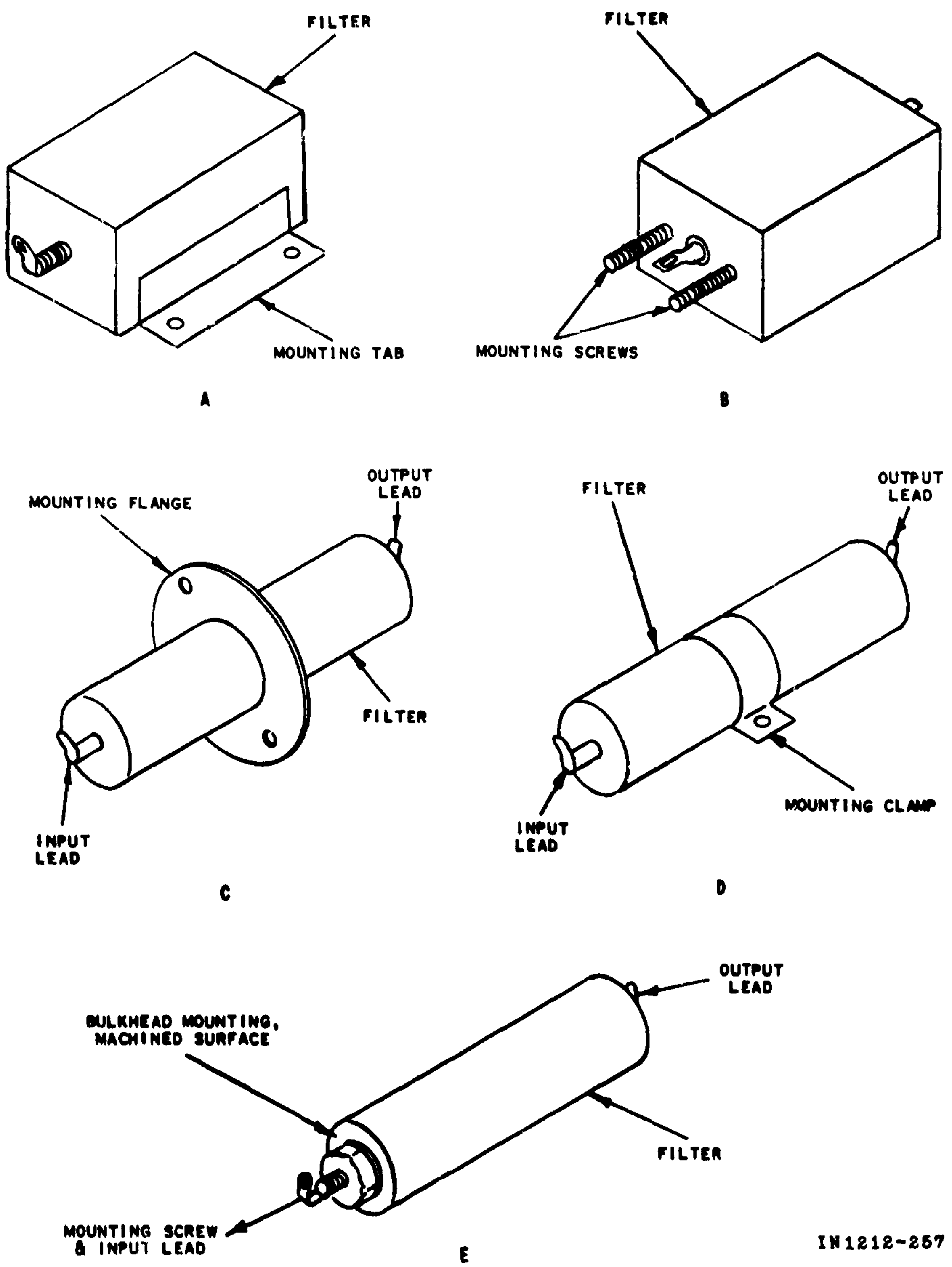


Figure 3-69. Typical Filter Mounting Techniques

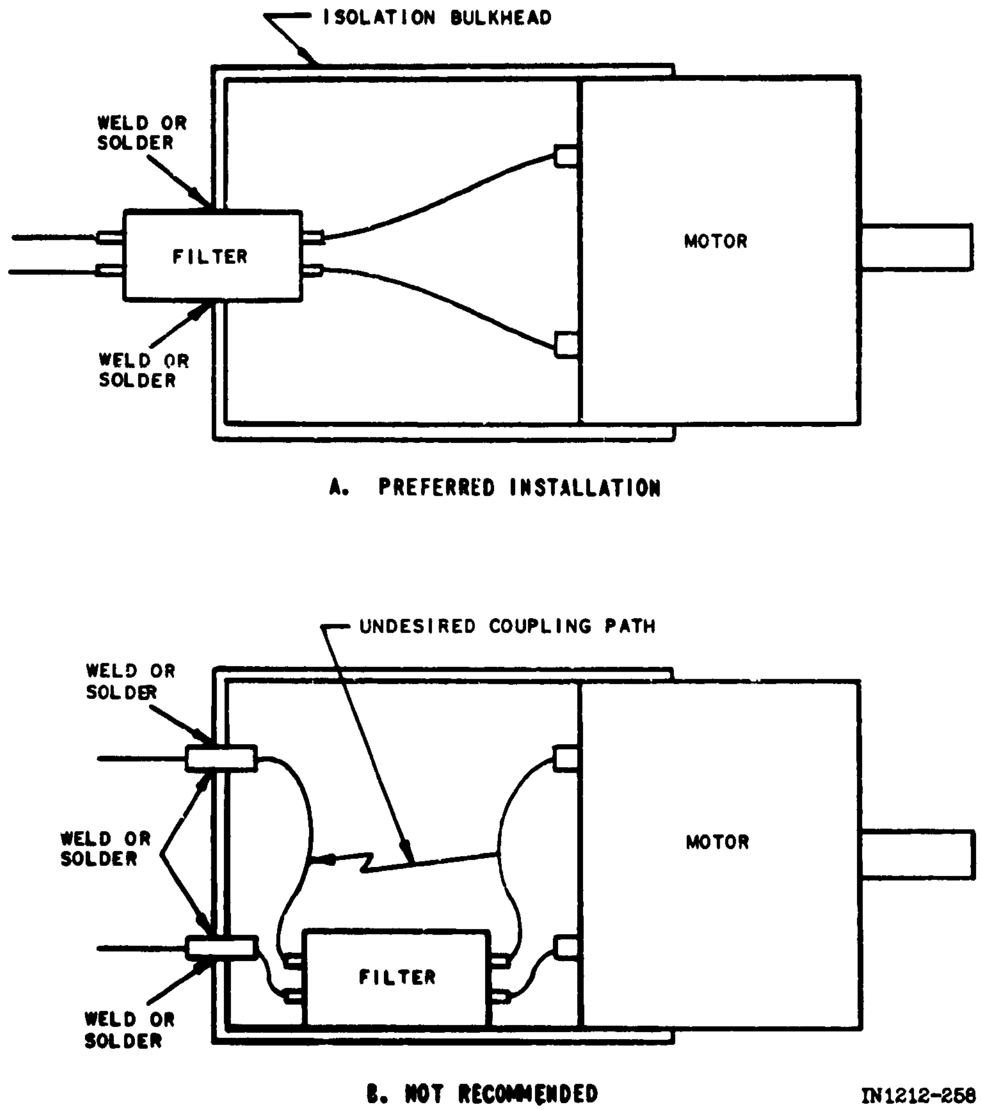


Figure 3-70. Filter Installation

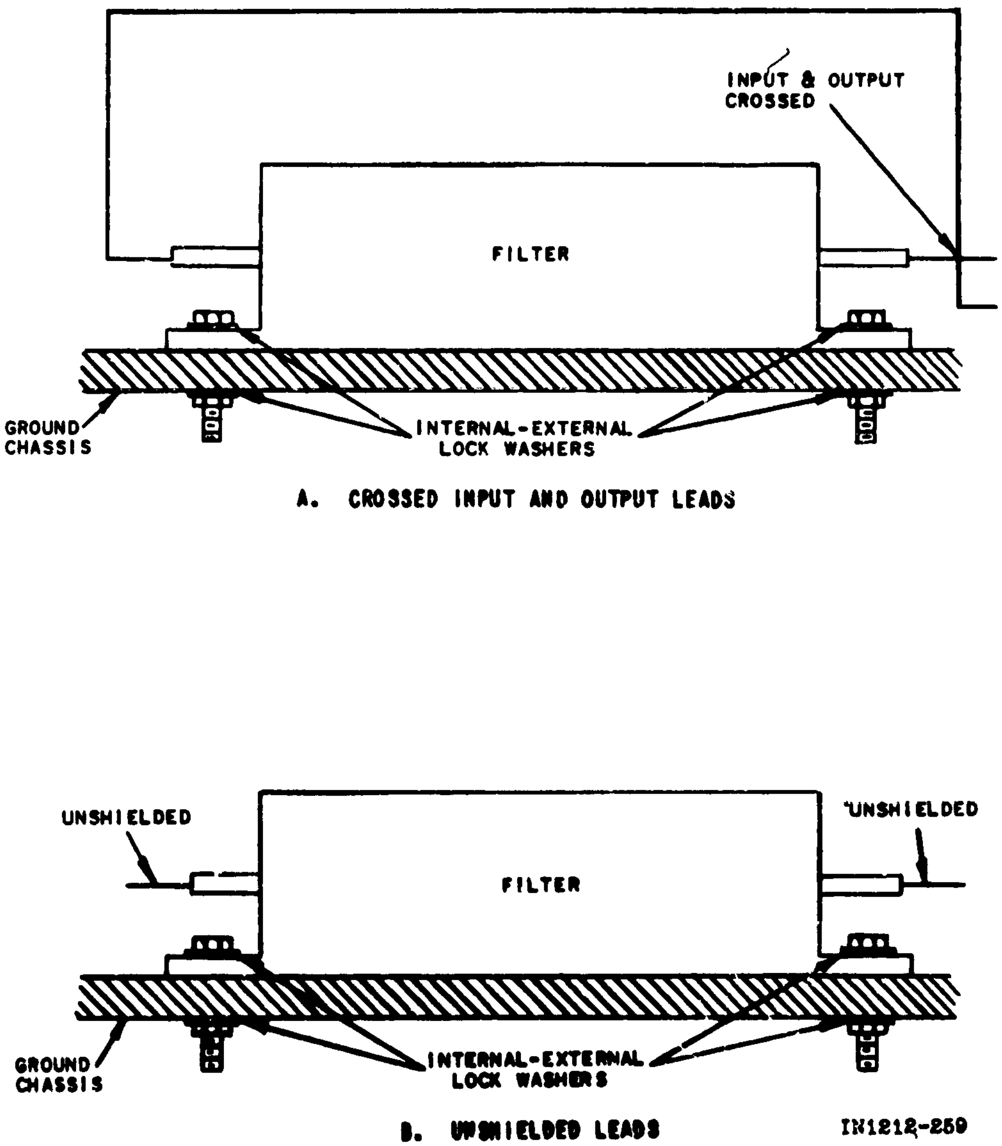


Figure 3-71. Incorrect Filter Mounting Methods

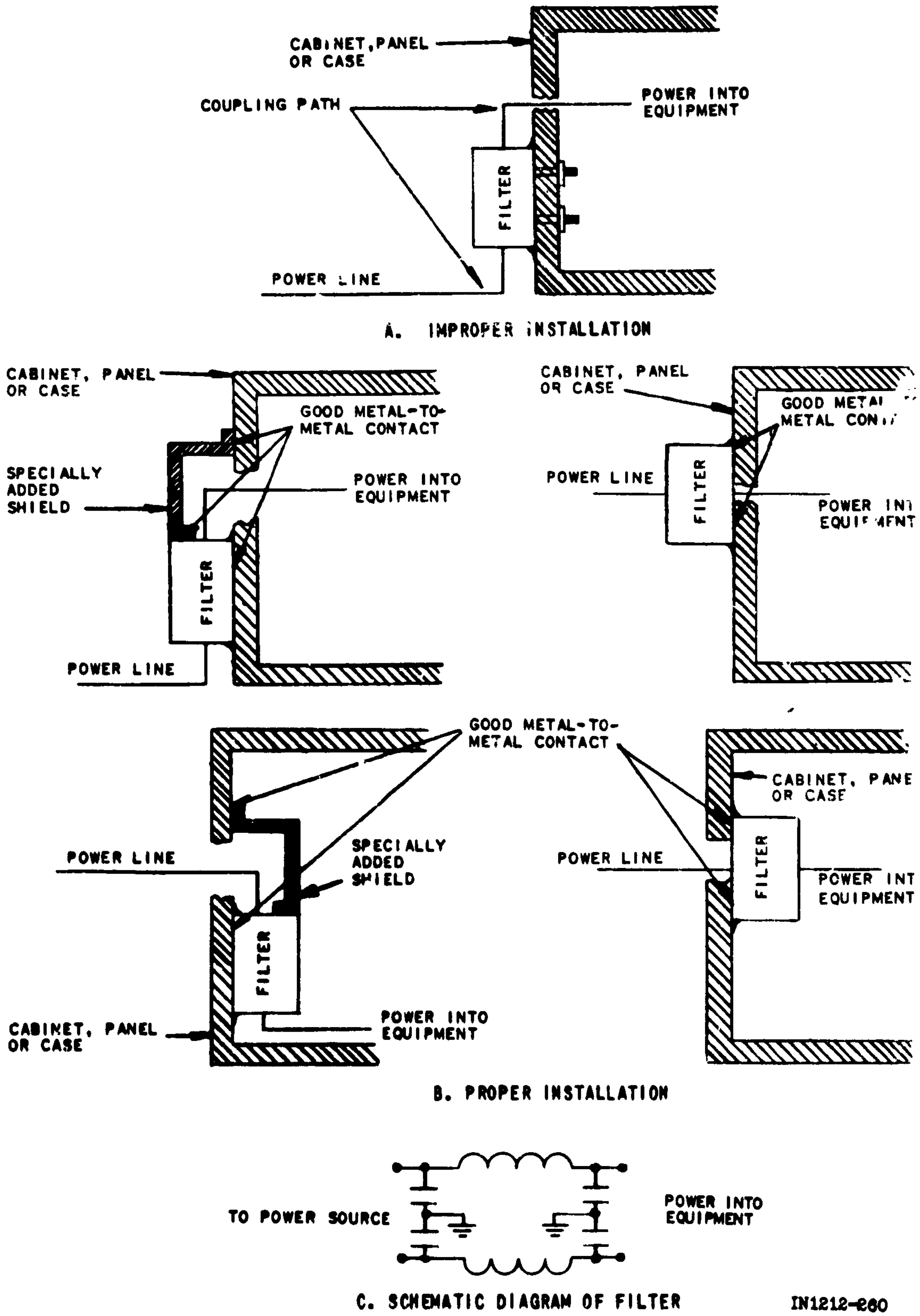


Figure 3-72. Installation of Power Line Filters

filter to the point of exit from the radiation area. The best way to effect a satisfactory filter installation is to specify feed-through or bulkhead mounting wherever feasible and to employ circumferential grounding of the filter-case to the bulkhead. Typical examples of these types of filter mountings are illustrated on figure 3-73. Figure 3-74 shows a feed-through capacitor installation in conjunction with shields. Feed-through capacitors, because of their superior characteristics, are recommended even when mounting in a shield is not feasible. Spot-facing of areas around filter mounting bolts is one reliable technique for obtaining an adequate bond. A far more permanent and better bond can be achieved by making the filter housing integral with the housing of the interference generator. The same bonding requirements apply when filters must be mounted on the susceptible unit rather than on the source.

3-12. Capacitor Selection

a. General

(1) The simplest and most common type of interference reduction network is a single capacitor used to shunt high-frequency interfering signals to ground. Such capacitors are widely employed in the power supply lines of electrical and electronic equipment, and may be used on both ac and dc leads. The optimum bypass capacitor is selected on the basis of its impedance-versus-frequency characteristic. The series resonant frequency of the optimum bypass capacitor should be approximately in the center of the interference frequency band to ensure that the bypass capacitor will have the lowest net impedance during operation of the circuit.

(2) The simple equivalent circuit for a fixed capacitor is a series RLC network, where R is due to dielectric losses and L is usually due to the inductance of the leads. Because lead inductance in a capacitor is a significant factor, it is possible to select a value of capacitance to create a series resonant circuit at the interference frequency. For a series resonant circuit, the net impedance includes

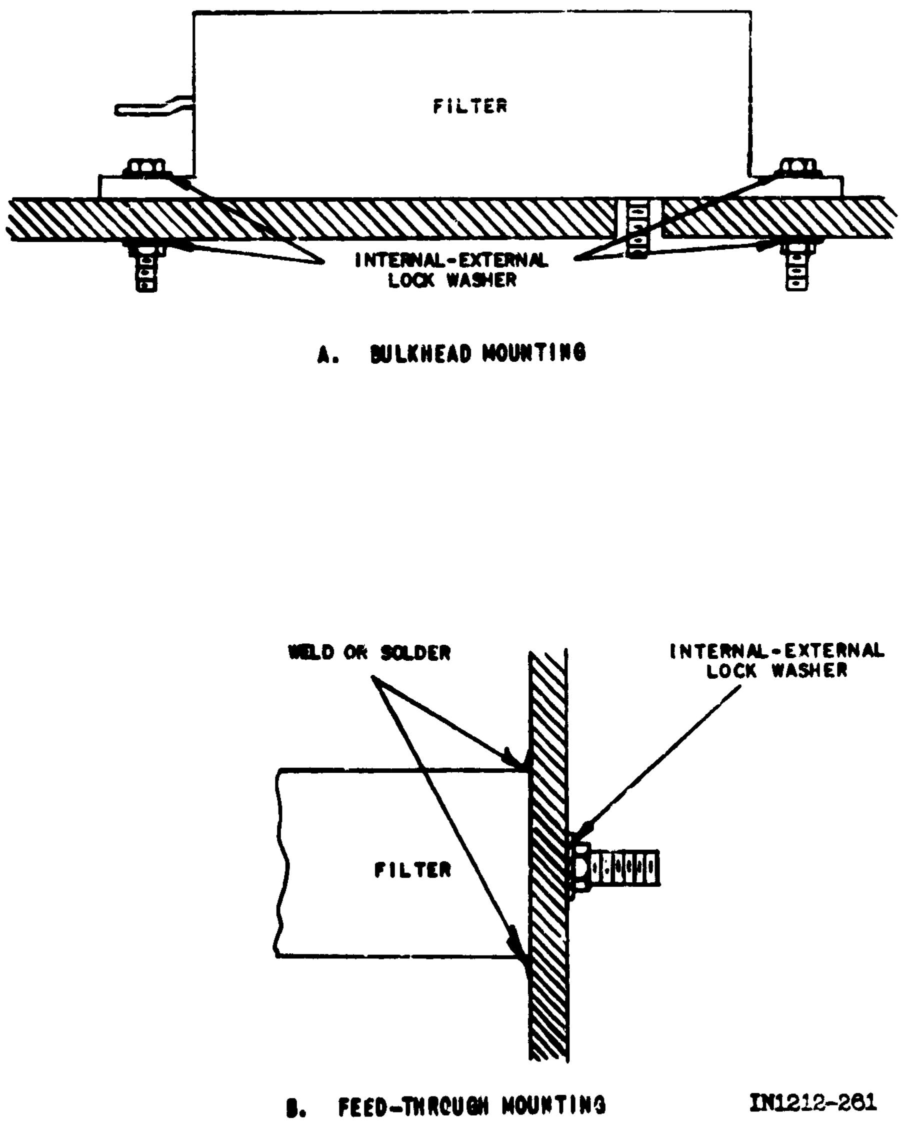


Figure 3-73. Typical Method for Mounting Filter Case

only the series loss resistance of the capacitor. With common mica or ceramic dielectric capacitors, the dissipation factor is very low. Thus, the equivalent series resistance is very small.

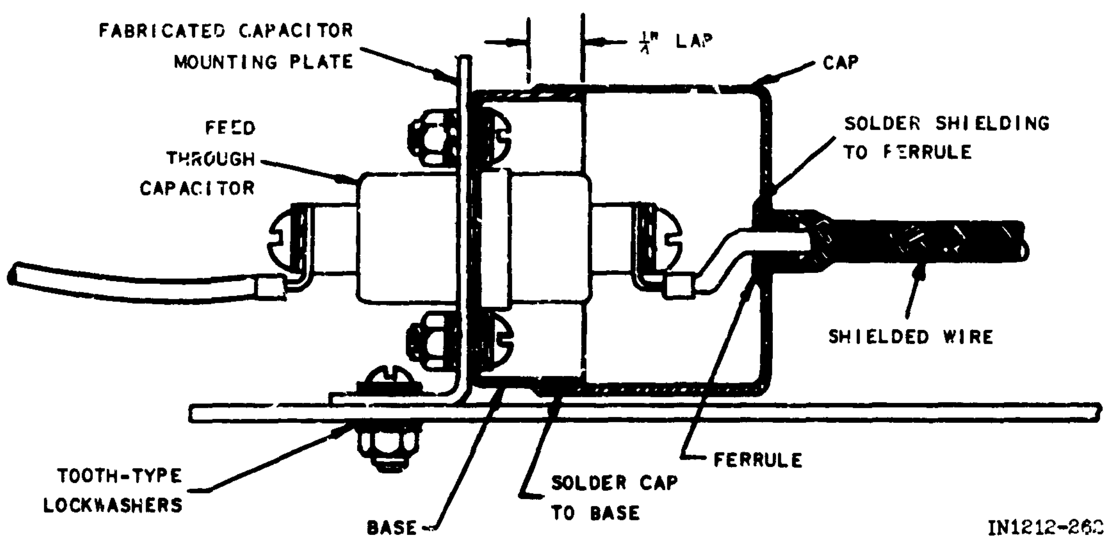


Figure 3-74. Feed-Through Capacitor Shield Assembly

(3) Consider the lead inductance and its effect upon the choice of capacitor values. The inductance of a straight piece of copper wire is computed as:

$$\text{Inductance} = 0.005L \left[2.3 \log_{10} \left(\frac{4L}{D} - 0.75 \right) \right] \mu h \quad (3\text{-}41)$$

where: L = wire length in inches

D = wire diameter in inches

The inductance of various lengths of AWG No. 12, 18, and 24 wire was calculated and the results plotted on figure 3-75. The reactance versus frequency curves for various values of lossless L and C elements is shown on figure 3-76.

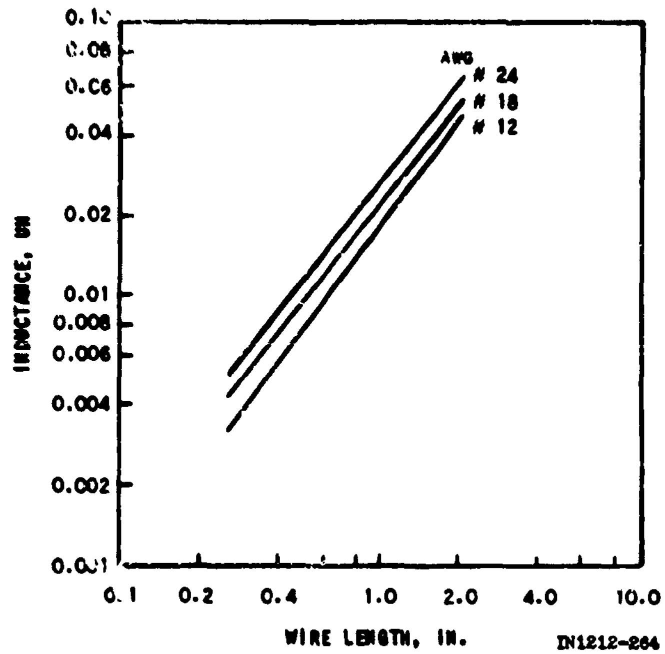


Figure 3-75. Inductance versus Length of Straight Round Wires

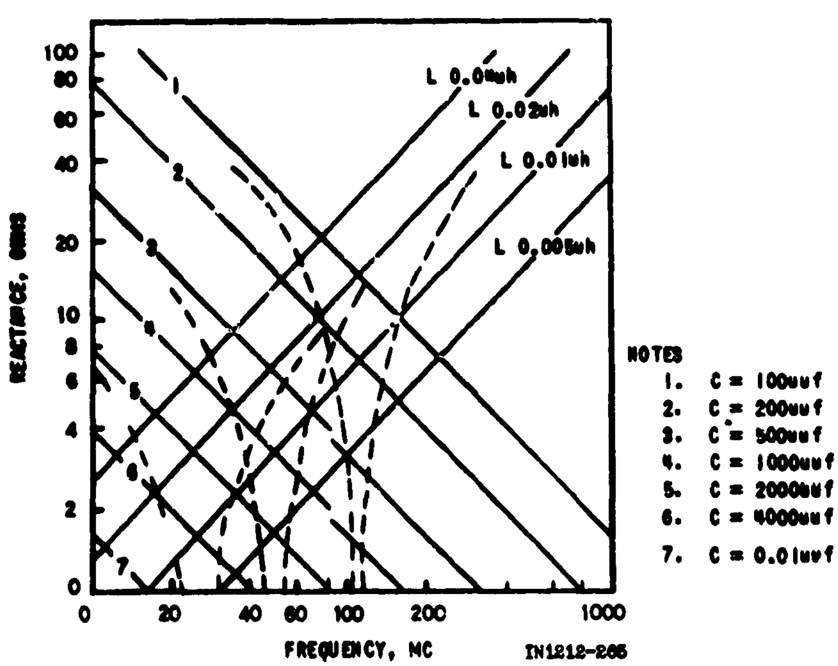


Figure 3-76. Reactance versus Frequency for Various Lossless L and C Elements

(4) The following examples show the significance of figures 3-75 and 3-76. Assume that a 0.44-inch length of No. 24 wire for each lead is acceptable. Since each lead has an inductance of 0.01 μh, the total series L in the equivalent circuit is 0.02 μh. Assume, also, that the optimum bypass capacitors for 25, 49, and 110 mc circuits are desired. For the stated L value, the optimum C values are 2000, 500, and 100 μμf, respectively, for the three frequencies, as shown on figure 3-76. Net impedance curves (assuming a zero dissipation factor) for each optimum capacitor are shown as dotted lines on figure 3-76.

(5) Consider the result of using too large a value of C. Assume that a 2000-μμf capacitor is incorrectly specified for a bypass capacitor in a 110-mc circuit. For the same total lead inductance, it is known from the previous example that this 2000-μμf capacitor has an inductive reactance of approximately 3.2 ohms at 25 mc and a reactance of 14.5 ohms at 110 mc. The optimum value for this 110-mc case is actually a 100-μμf capacitor (fig. 3-76). This capacitor produces a net impedance of 1 ohm or less at frequencies from 107 to 115 mc, and a net impedance of 10 ohms or less from 80 to 155 mc.

(6) To measure the approximate value of f_o, the series resonant frequency, connect the capacitor leads together (using the same lead length as required in the circuit) and couple this one-turn loop into a grid-dip meter. The resultant f_o will not be exactly the series resonant frequency for the same capacitor when soldered in a circuit because the L of a small one-turn loop is not the same as the L of two small straight lengths of wire, but this measured value of f_o is usually within 10 per-cent of the true value.

(7) The effectiveness of a capacitor as an interference reduction component may be measured in terms of its insertion loss. This loss, as an index of effectiveness for capacitors, may be expressed as the ratio of the interference voltage before and after insertion of the capacitor. Insertion loss data yields a qualitative measure of the interference reduction that may be effected at a load by the application of a particular capacitor.

b. Lead-Type Capacitors.

(1) The impedance of an ideal capacitor follows the relationship: $X_C = \frac{1}{\omega C}$. As previously stated, a practical capacitor possesses inductance and resistance in series with its capacitance. The inductance consists of the inductance inherent in the capacitor itself and the inductance of the capacitor leads. To reduce the internal inductance, non-inductive capacitors are constructed by winding two layers of aluminum foil, separated by a dielectric material such as kraft paper or mylar, into a cylindrical roll (fig. 3-77). One foil projects beyond the dielectric on each end. The ends are then swaged by pressing against a wheel that revolves through a reservior of molten aluminum. This action effectively bonds every turn of each foil together, resulting in very low inductance between the windings. When $X_L = X_C$, the capacitor becomes a series resonant circuit with very usable rf characteristics. At the resonance point, the impedance is a minimum and the insertion loss is at its maximum. Above the resonant frequency, ωL predominates, and the capacitor ceases to be an effective by-pass element.

(2) For the circuit of figure 3-77, the resonant frequency is given by the expression $f_r = \frac{1}{2\pi\sqrt{LC}}$, where L is in henries and C is in farads. The resonant frequency increases as the total inductance decreases (the smaller the total inductance, the higher the resonant frequency and the greater

the useful frequency range). For a given capacitor, the internal inductance is fixed, but the external inductance is a function of lead length, as shown by equation 3-41.

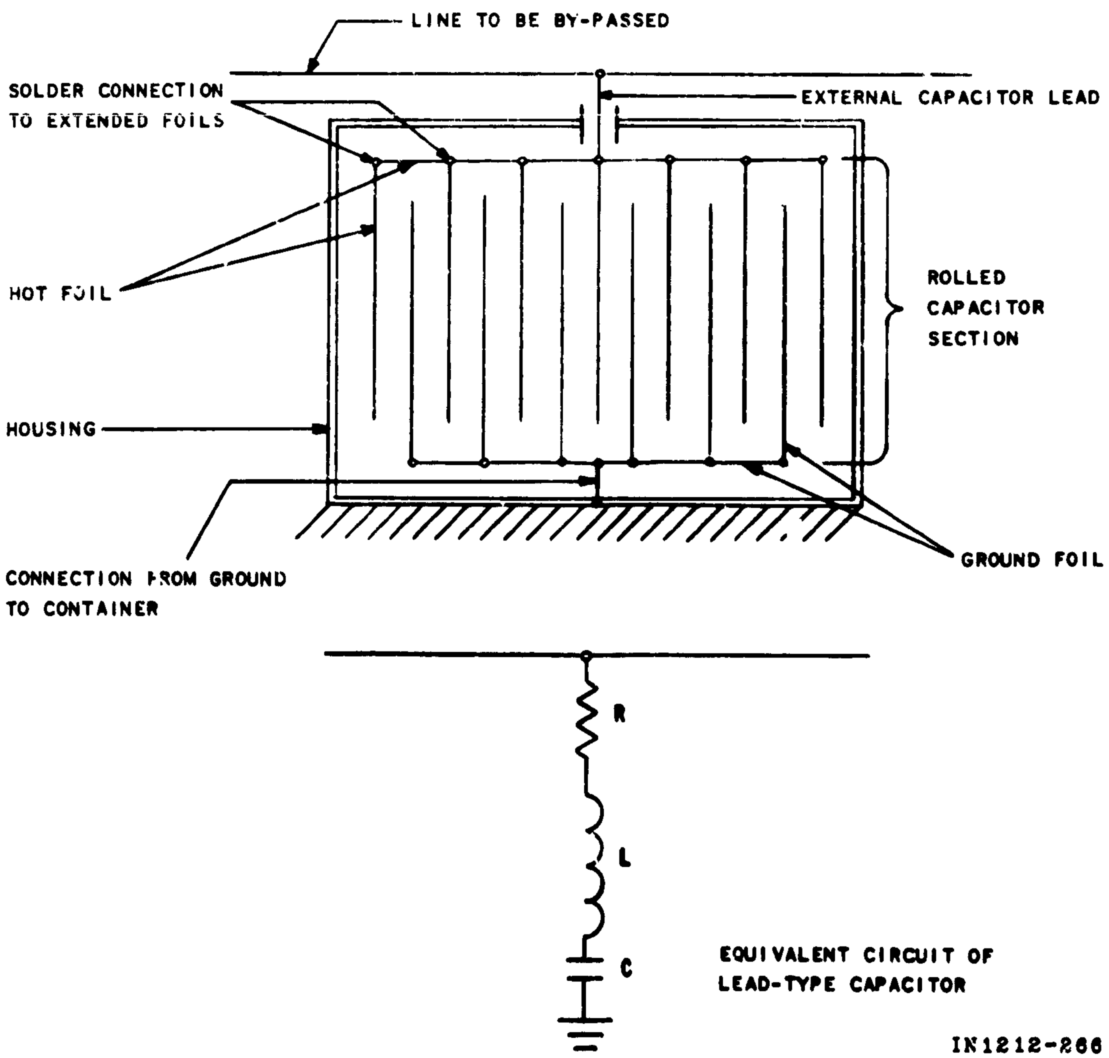


Figure 3-77. Typical Bypass Capacitor Construction with Equivalent Circuit

(3) Varying the lead length of a capacitor and, consequently, its inductance, offers some degree of control in shifting its resonant frequency. Figures 3-78 through 3-81 provide a series of design curves relating lead length to resonant frequency for several types of commercially available capacitors. The paper tubular type is not acceptable for rfi suppression use. The flexibility of capacitor values using lead length as the variable parameter is easily demonstrated: Suppose the requirement is a by-pass capacitor application at a frequency of 2 mc. Using paper tubular capacitors, whose resonant frequency characterisitics are given in figure 3-78, the following choices are available:

1) 0.5 μf with 0.15-inch leads

2) 0.25 μf with 0.5-inch leads

3) 0.1 μf with 1.5-inch leads

4) 0.05 μf with 2.6-inch leads

From these design curves, and from figure 3-82 it may be seen that:

1) The lead-type capacitor, near resonance, is somewhat superior to the ideal capacitor of similar value which by-passes without resonance by capacitive reactance alone

2) Twenty percent beyond the resonant frequency, the lead-type capacitor is ineffective as an interference reduction component

3) Even by decreasing the external lead length as much as reasonably possible, the lead-type capacitor has a frequency limit imposed on it by its internal inductance

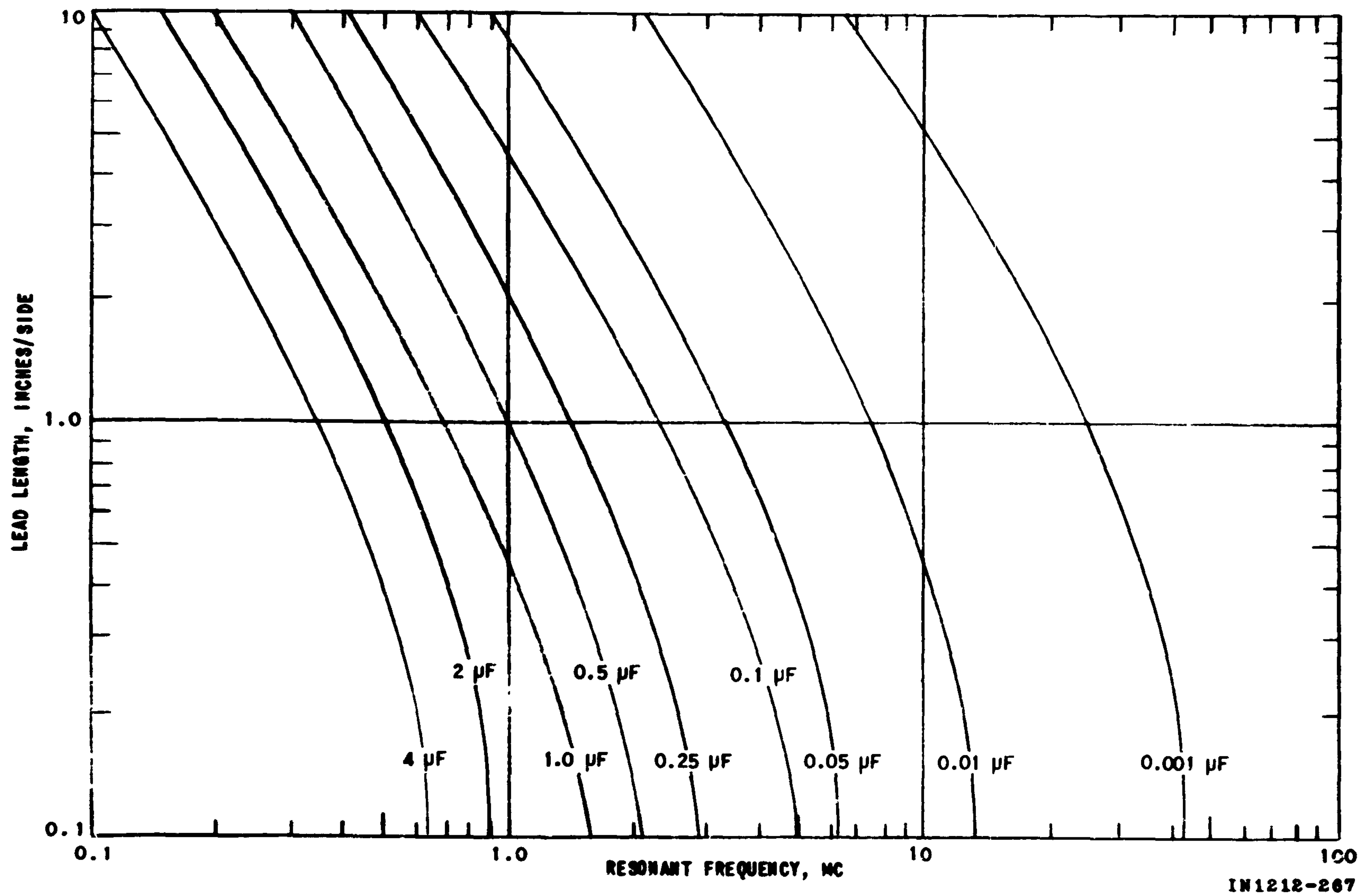
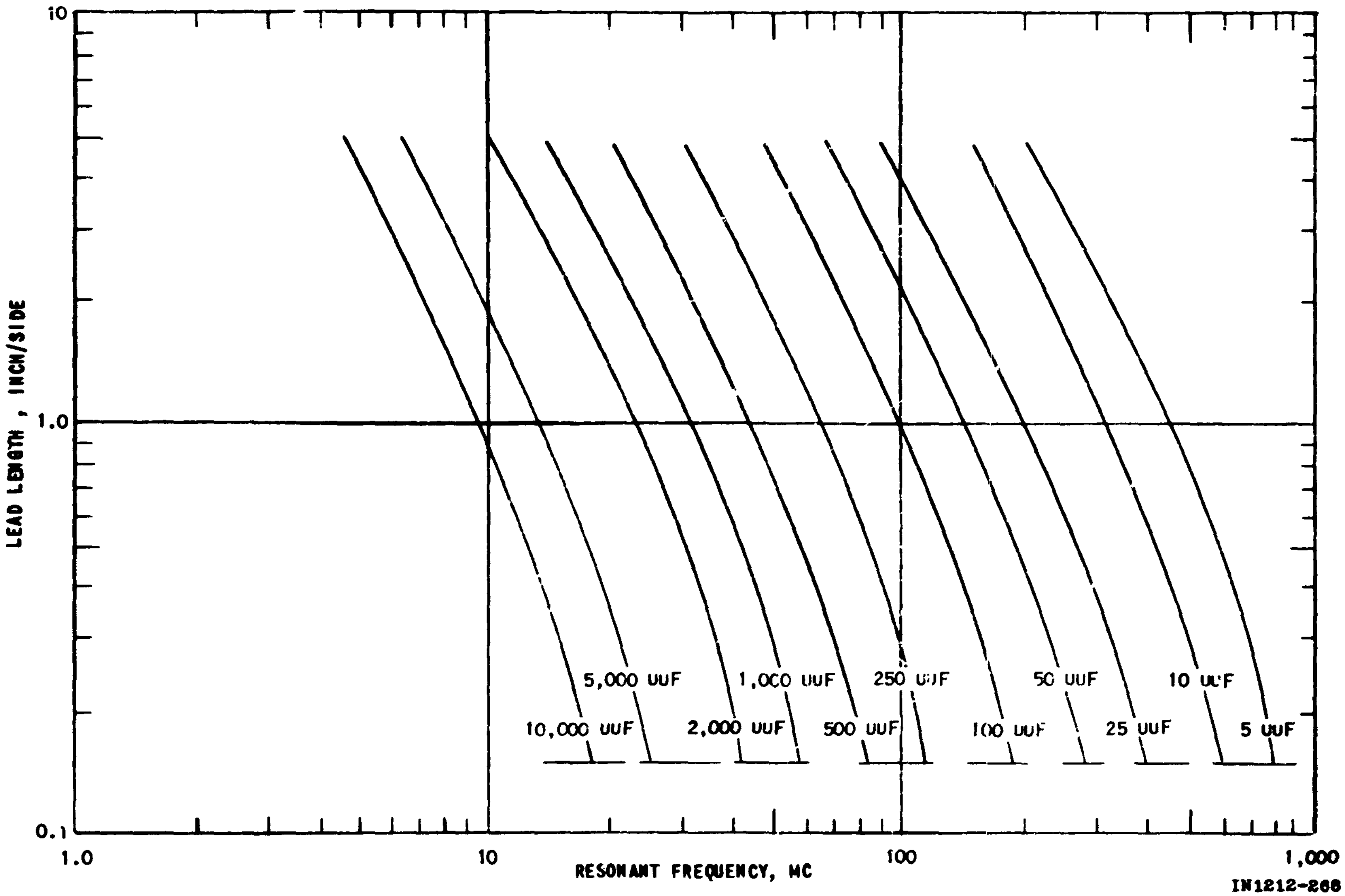


Figure 3-78. Resonant Frequency as a Function of Lead Length for Paper Tubular Capacitors



[illegible] Frequency [illegible] Lead Lengt[illegible] Mica Capacitors

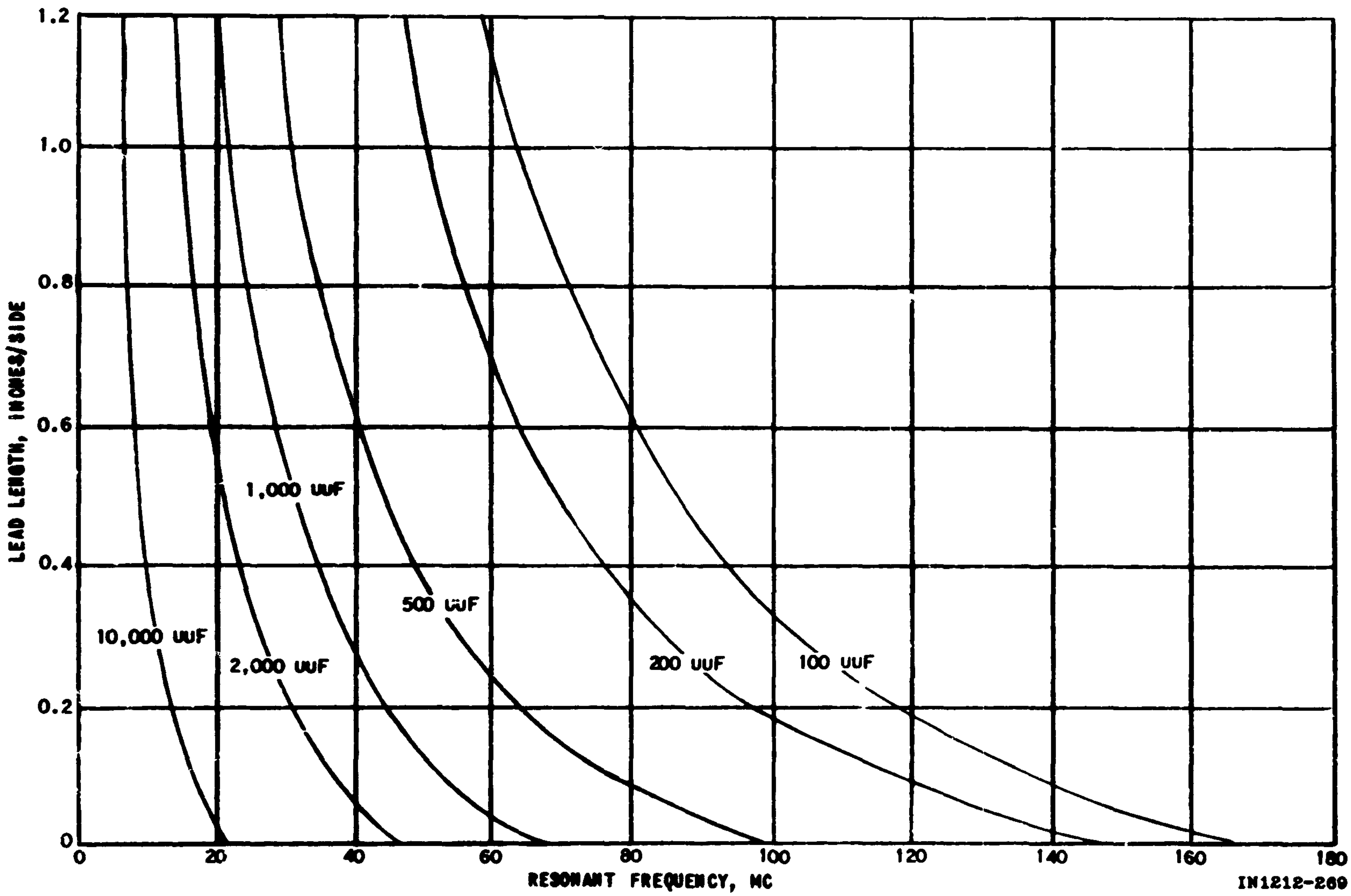


Figure 3-80. Resonant Frequency as a Function of Lead Length for Disc Ceramic Capacitors

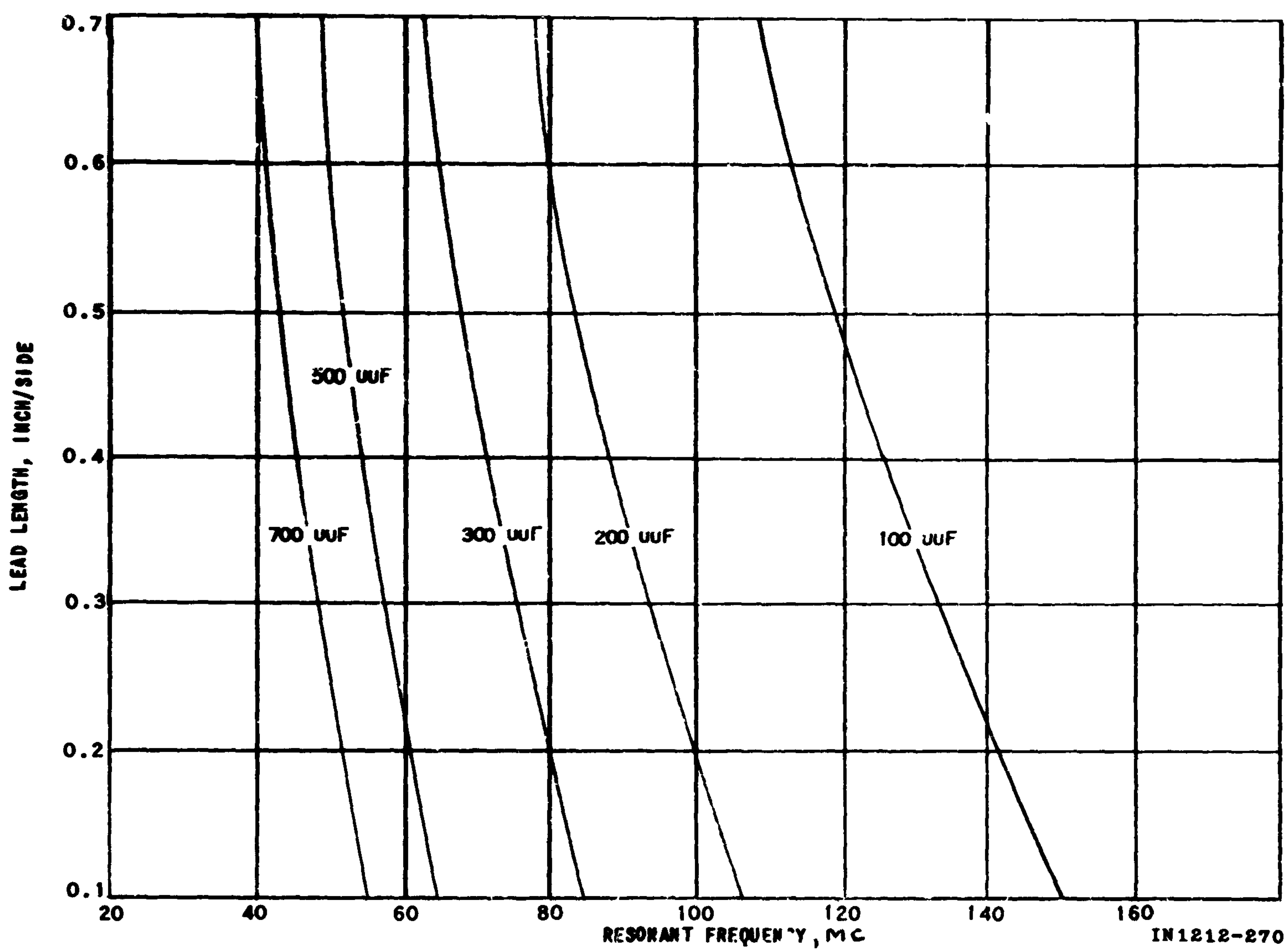


Figure 3-81. Resonant Frequency as a Function of Lead Length for Standoff Type Ceramic

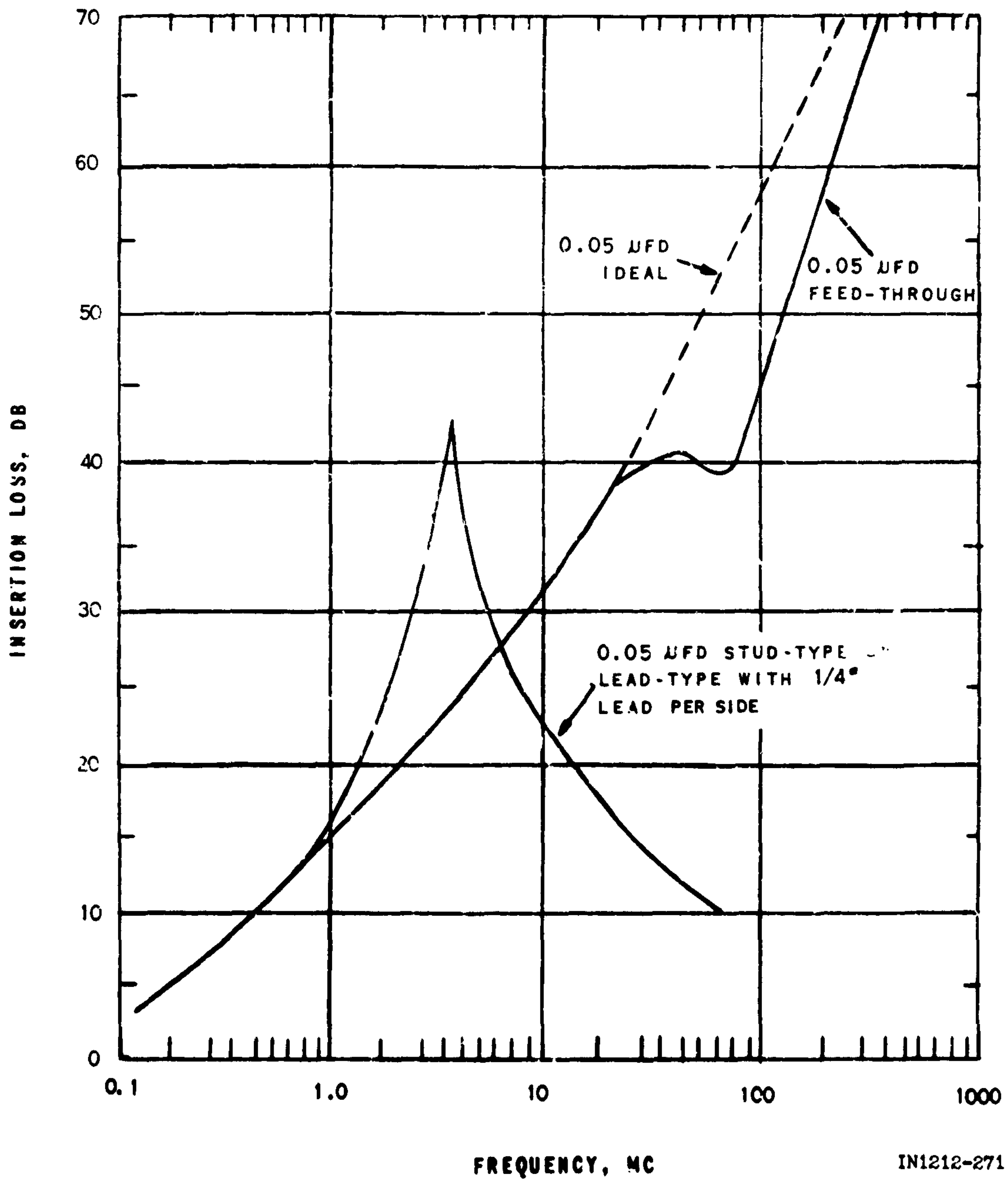


Figure 3-82. Insertion Loss Comparing Ideal, Feed-Through and Lead-Type Capacitors for a 50-Ohm System

The resonant frequencies are only relative since they vary with the geometry and construction of the component. In practice, the resonant frequency is checked with a grid-dip meter. The ceramic standoff, or stud-type capacitor, represents an effort to reduce the inductance and raise the resonant frequency. It is still basically a lead type unit, with one relatively long pigtail lead replaced by a short, thick stud terminal having a minimum of inductance. Although this terminal improves insertion-loss characteristics, the resonant frequency of the stud-type capacitor is still only slightly higher than that of a conventional lead-type capacitor that has a short lead comparable to the length of the stud.

c. Feed-Through Capacitors.

(1) In those cases where interference suppression requires broadband characteristics that the ordinary capacitor does not provide, two components may be used; these are the feed-through capacitor and the low-pass filter. A typical feed-through capacitor is shown on figure 3-83. It features reduced internal inductance and freedom from the external inductance that is common to all lead-type capacitors. The feed-through capacitor may be considered as a four-terminal network, similar to a section of coaxial line. The input and output terminals are connected; and capacitance exists between both of these terminals and the case of the capacitor.

(2) To maintain the highest possible resonant frequency, it is essential that connections within the capacitor be as short as possible. Thus, a good feed-through capacitor has soldered internal connections so that internal inductance and resistance are at their absolute minimum. In addition, the feed-through principle acts to cancel

out any induced fields resulting from current flow in the foil itself. At present, feed-through capacitors that resonate above 1000 mc can be produced.

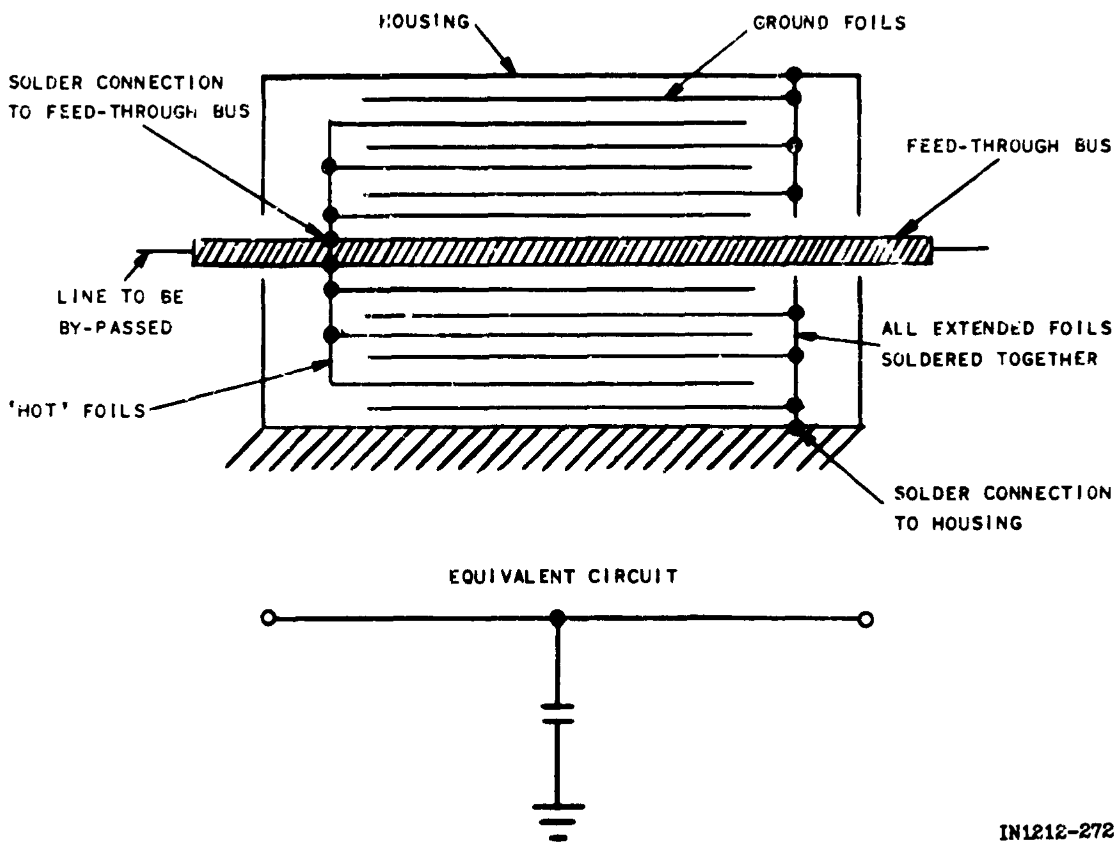


Figure 3-83. Typical Feed-Through Capacitor Construction and Equivalent Circuit

(3) The attenuation characteristic of a feed-through capacitor is simliar to that of any low-pass filter with the exception that much more capacitance is required to obtain a given insertion loss at the cutoff frequency. To realize the fullest advantage of the feed-through capacitor as a filter, it is most important that it be mounted so that the input is completely shielded from the output. Bulkhead mounting is one method commonly employed. The insertion loss will then very closely fol ow the characteristic of an ideal capacitor. When the feed-through capacitor is

inserted into a matched 50-ohm line, the insertion loss can be computed as follows:

$$\text{Insertion loss in db} = \alpha = 10 \log_{10} \left[1 + \left(\frac{Z_o \omega C}{2} \right)^2 \right]$$

$$\alpha = 20 \log_{10} (50\pi fC) \qquad fC \geq \text{approx } 1/15 \qquad (3\text{-}42)$$

$$= 44 \text{ db} + 20 \log_{10} (fC)$$

This equation is valid when the insertion loss is significantly greater than zero. Typical curves of insertion loss, based on a 50-ohm line for various size feed-through capacitors, are shown on figure 3-84. Figure 3-82 is a curve of insertion loss of a feed-through capacitor compared to an ideal capacitor and a lead-type capacitor for a 50-ohm system. The required capacitance value to yield a given insertion loss at a given frequency can be read directly from the curves of figure 3-84. Alternately, it may be computed from the following equation:

$$C = 1/f \text{ Antilog}_{10} \frac{\alpha - 44}{20} \qquad (3\text{-}43)$$

where: C = capacitance in μf (for a 50-ohm system)

f = frequency in mc

α = insertion loss in db

If the frequency of interest falls within the range where the characteristic departs from that of the ideal capacitor, a safety factor of about 10 db should be added to the required α. This also applies in the application of the curves of figure 3-84, which are based on ideal characteristics.

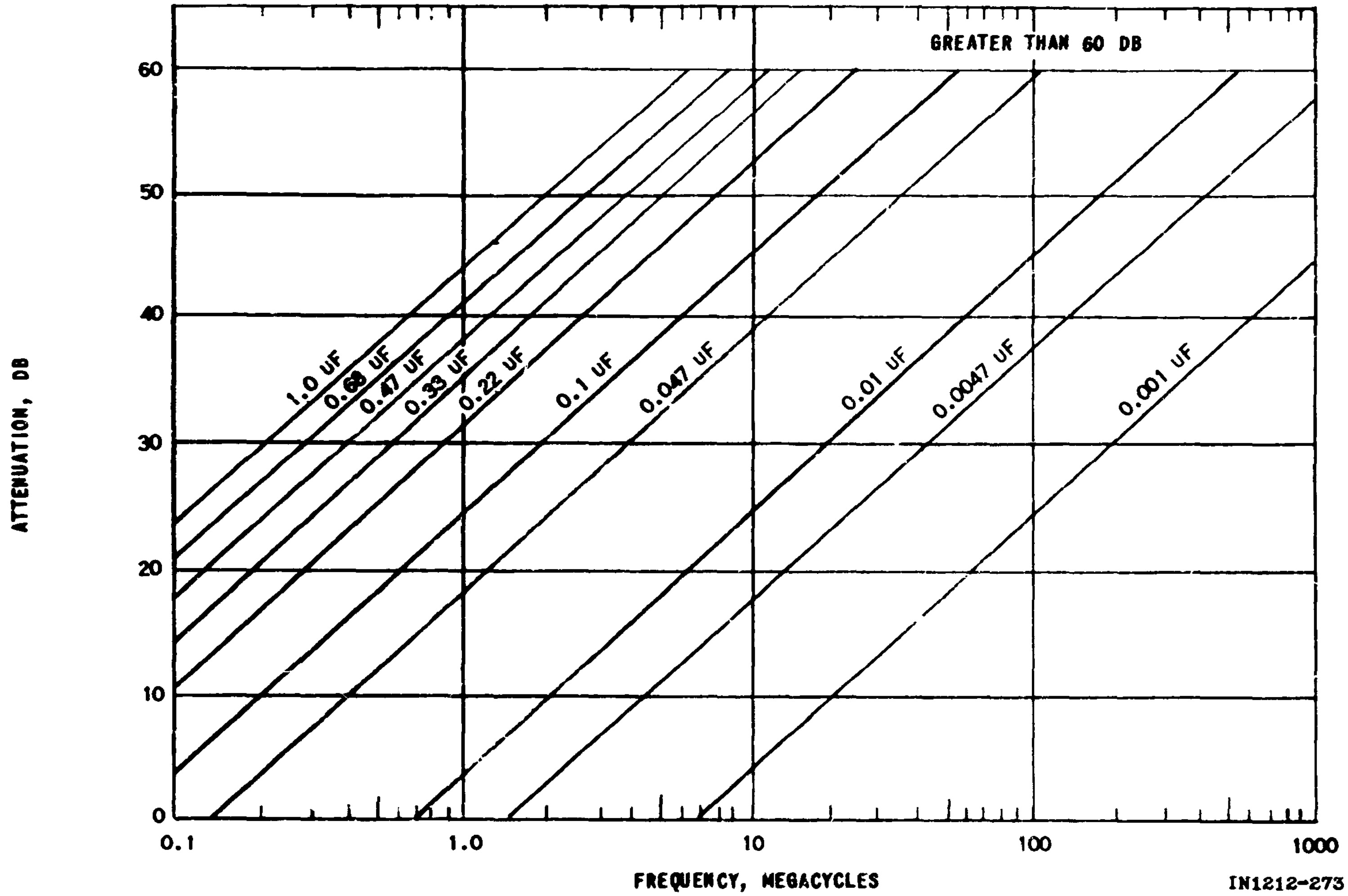


Figure 3-84. Typical Ideal Insertion Losses of Feed-Through Capacitors in a 50-Ohm Line

(4) Selection of the optimum feed-through capacitance value usuall necessitates a compromise between a capacitance sufficiently large to provide good insertion loss at the frequency of concern, yet not so large as to cause appreciable loading at 60 cps or at other power frequencies. Applying capacitanc across switches necessitates care, as capacity in excess of 0.01 μf may cause excessive metal transfer or burning and sticking of the contacts. While the specific application will determine the choice of capacitor, for most design purposes, the range of capacitance values is from 0.01 to 2 μf. In most applications, unless the circuit contains high levels of interference voltages at low frequencies, a feed-through capacitor may be used instead of a filter net work to provide the necessary by-passing effectiveness.

(5) Feed-through capacitors must carry all the line current through their center conductor. They must, therefore, be rated in terms of current-carrying capacity as well as capacitance and voltage. New low-voltage, high-capacity, small-sized, feed-through electrolytic capacitors provide effective reduction of interference from audio through uhf frequencies. Their insertion-loss characteristics are simila to those of comparably-rated paper feed-through capacitors, though their physical size is much smaller.

3-13. Inductors

Inductors are often used for interference reduction design applica... The voltage across an inductor is $-L\frac{di}{dt}$; that is, the induced electro- motive force opposes any variation or interruption of the applied current. When inductors are used as interference reduction devices, they must be capable of passing the operating current without excessive heating and without causing effects upon, or being affected by, nearby electric or magnetic fields. They must also preserve their electrical properties over as wide a frequency range as possible.

a. Coils. Figure 3-85 shows some typical coil configurations. The windings may be either solenoid (cylindrical) as in fig. 3-85 A, B, C, D, E, or toroid (circular) as in F and G. The cores may be air, powdered iron, ferrite, or molybdenum permalloy. The reactance of a pure inductor increases directly with frequency and with the value of the inductor. Because of the increase of inductive reactance with an increase in frequency, a series-connected inductance acts similarly to a low-pass filter, and a parallel-connected inductance acts similarly to a high-pass filter. At low frequencies, the value of the inductance must be larger than it would need be to give the same impedance at a higher frequency. The following are coil characteristics:

1) Inductance
2) Permeability
3) $Q = X_L/R_E$
4) Resistance, $R_E = R_{dc}$ + field loss
5) Heat rise
6) Field losses
7) Coupling within the inductive field
8) Saturation (excessive flux density)
9) Distributed capacitance

Figure 3-86 shows the equivalent circuit of an inductance coil in which: L = inductance, C = distributed capacitance, R_E = winding resistance, and R_s = shunting effect of the losses in the surrounding medium. Because of the way that coils are wound, there is some distributed capacity between windings. This capacity has an important effect upon the action of the coil at high frequencies. At a frequency determined by its inductance and distributed capacity, the coil becomes a parallel resonant circuit. At frequencies below this resonance, the coil is predominantly inductive; at frequencies above resonance, the coil is predominantly capacitive. The distributed capacity of a coil constitutes the limiting factor on the values of inductance that may be used in filters.

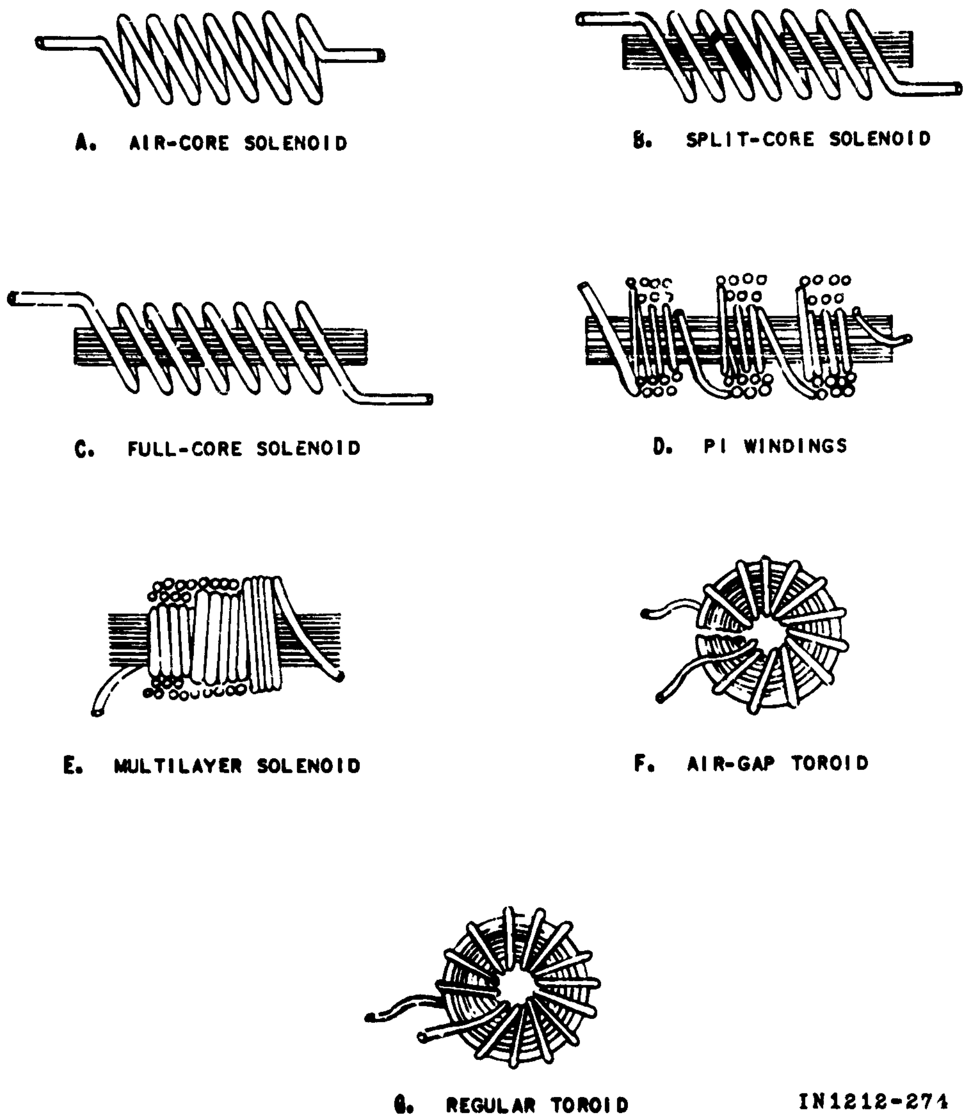


Figure 3-85: Typical Coil Configurations

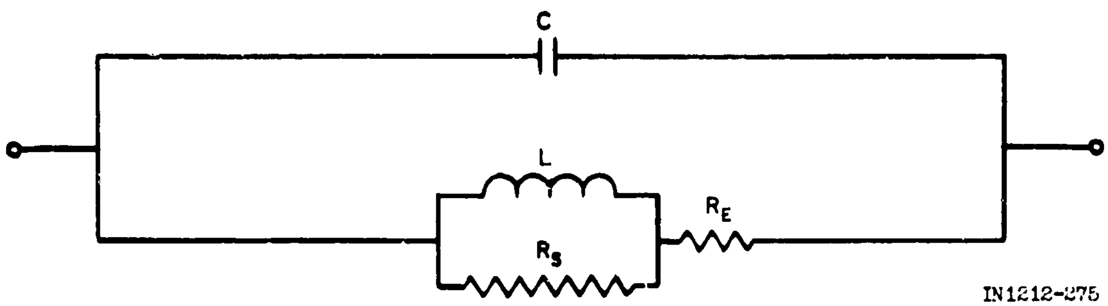


Figure 3-86. Equivalent Circuit of an Inductance Coil

If an attempt is made to use too large an inductance, the distributed capacity may cause the coil to resonate at too low a frequency. The distributed capacity of coils may be reduced by winding the coil in several sections so that the distributed capacities of the various sections are in series with each other. Thus, where high insertion loss is desired at low frequencies, the use of small inductances with additional filter sections is frequently a better solution than the use of a single-section filter. The distributed capacity of a single-layer coil can be found by the use of the nomograph of figure 3-87 once the the diameter of the coil (d), the diameter of the wire used (a), and the distance between centers of adjacent turns (s) are known.

b. Toroids. When greater inductance is required in a given space than is obtainable with an air core, a magnetic core material is used. When magnetic core materials are used in filters, the insertion loss of the filter must be specified as the insertion loss under load because magnetic core materials without external air gaps tend to saturate when used in dc circuits drawing large amounts of current. In such a case, the insertion loss changes appreciably. The use of nonsaturable materials for the inductance core is recommended. Often, inductance core material is specified which, under no load, gives good insertion-loss characteristics, but under full load saturates, causing lowered inductance and reduced insertion loss. The effect

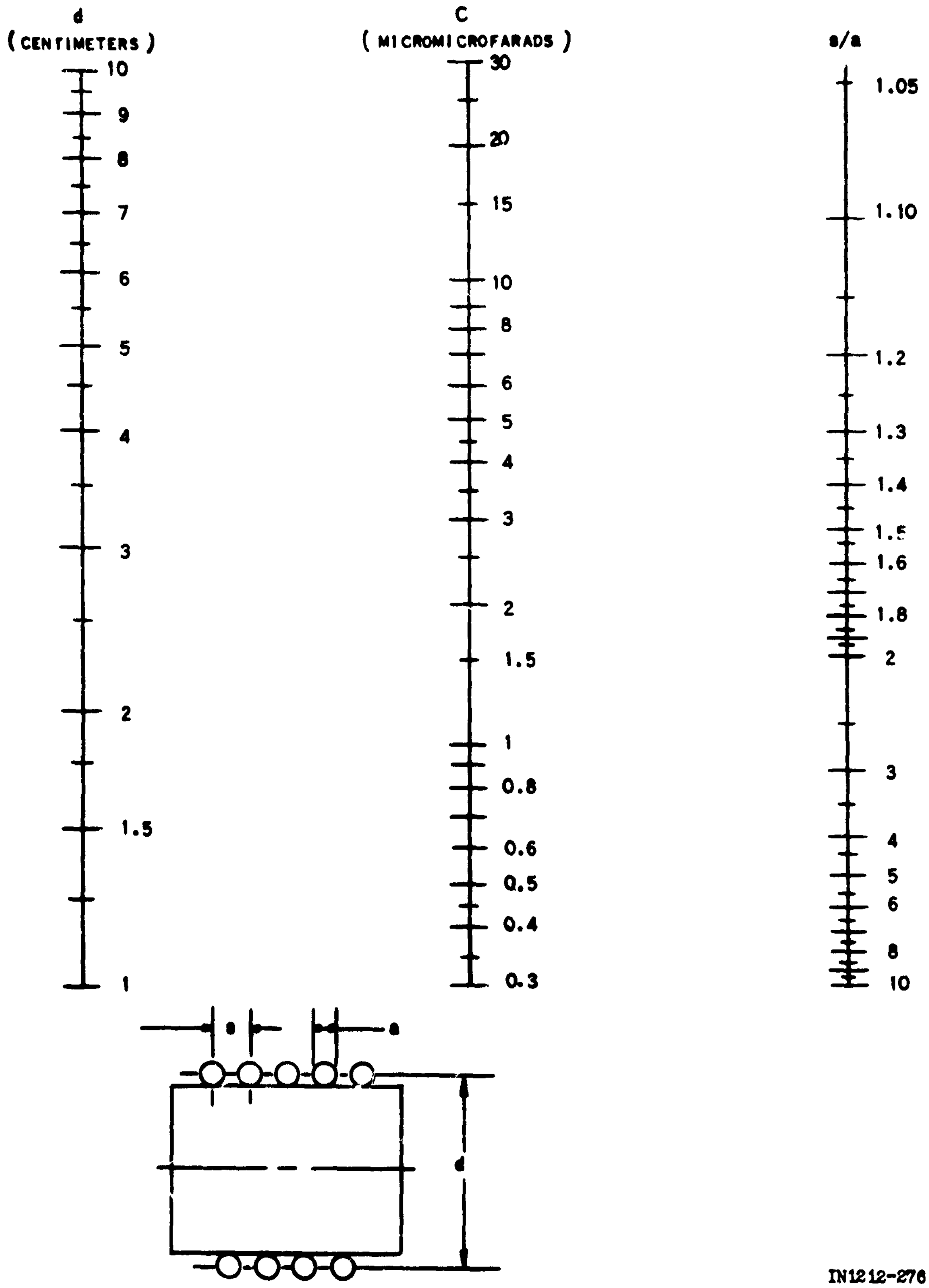


Figure 3-87. Nomograph for Determining the Distributed Capacitance of Single-Layer Coils

of saturation or the inductance is evident from the inductance equation of a toroid: For a single-layer toroid of rectangular cross-section, with or without intrinsic air gap, but without external air gap (fig. 3-88):

$$L = 0.00508\ N^2\ bu_d\ \ln \frac{r_2}{r_1} \qquad (3\text{-}44)$$

For a single-layer toroid of round cross section (fig. 3-89):

$$L = 0.0319N^2 u_d \left(r_m - \sqrt{r_m^{\ 2} - b^2/4} \right) \qquad (3\text{-}45)$$

where: L = inductance, in μh

b = core width, in inches

u_d = average incremental permeability

r_1 = inside radius, in inches

r_2 = outside radius, in inches

r_m = mean radius, in inches

N = total number of turns

In both cases (round and rectangular cross section), the inductance is directly proportional to the average incremental permeability. The permeability is a function of the magnetization which varies as:

$$H = \frac{NI}{2\pi\ r_m} = \frac{NI}{\pi\ (r_1 + r_2)} \qquad (3\text{-}46)$$

where: H = magnetization, in ampere-turns per inch

I = dc current, in amperes

Figure 3-90 shows the variation of incremental permeability with magnetization for different materials. The ferrites have the highest initial permeability, but they reach saturation very quickly. Also, they have the lowest losses at 200 kc, but their losses increase very rapidly with increasing frequency. Molypermalloys do

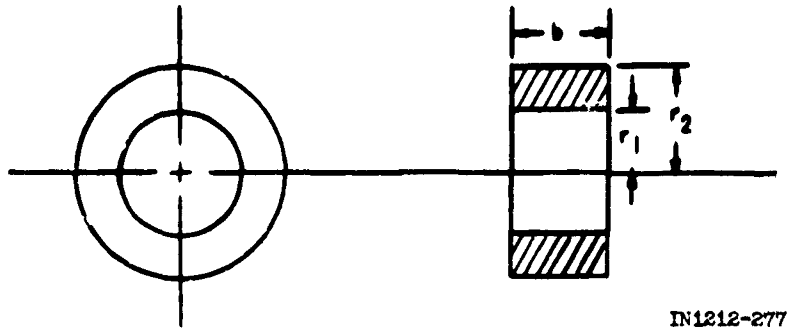


Figure 3-88. Toroid of Rectangular Cross-Section

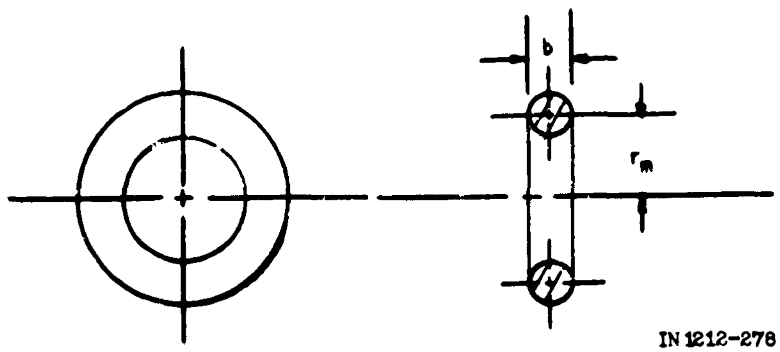


Figure 3-89. Toroid of Round Cross-Section

not have very high initial permeabilities; however, they do not saturate easily and their losses at high frequencies are low. The molypermalloys at the present time appear to be the best compromise available for those cases where high inductances are needed with low losses at the high frequencies. It should also be determined that temperature changes that may be encountered do not change the core inductance sufficiently to modify the filter's characteristics.

3-14. Faraday Shields in Low-Level Transformer Design

a. General. Low-level transformers are often subject to interference voltages. The cause of these voltages could be stray magnetic fields, common-mode signals, or machine-made disturbances. An understanding by the design engineer of shielding techniques as used in the construction of low-level input transformers, chopper input transformer

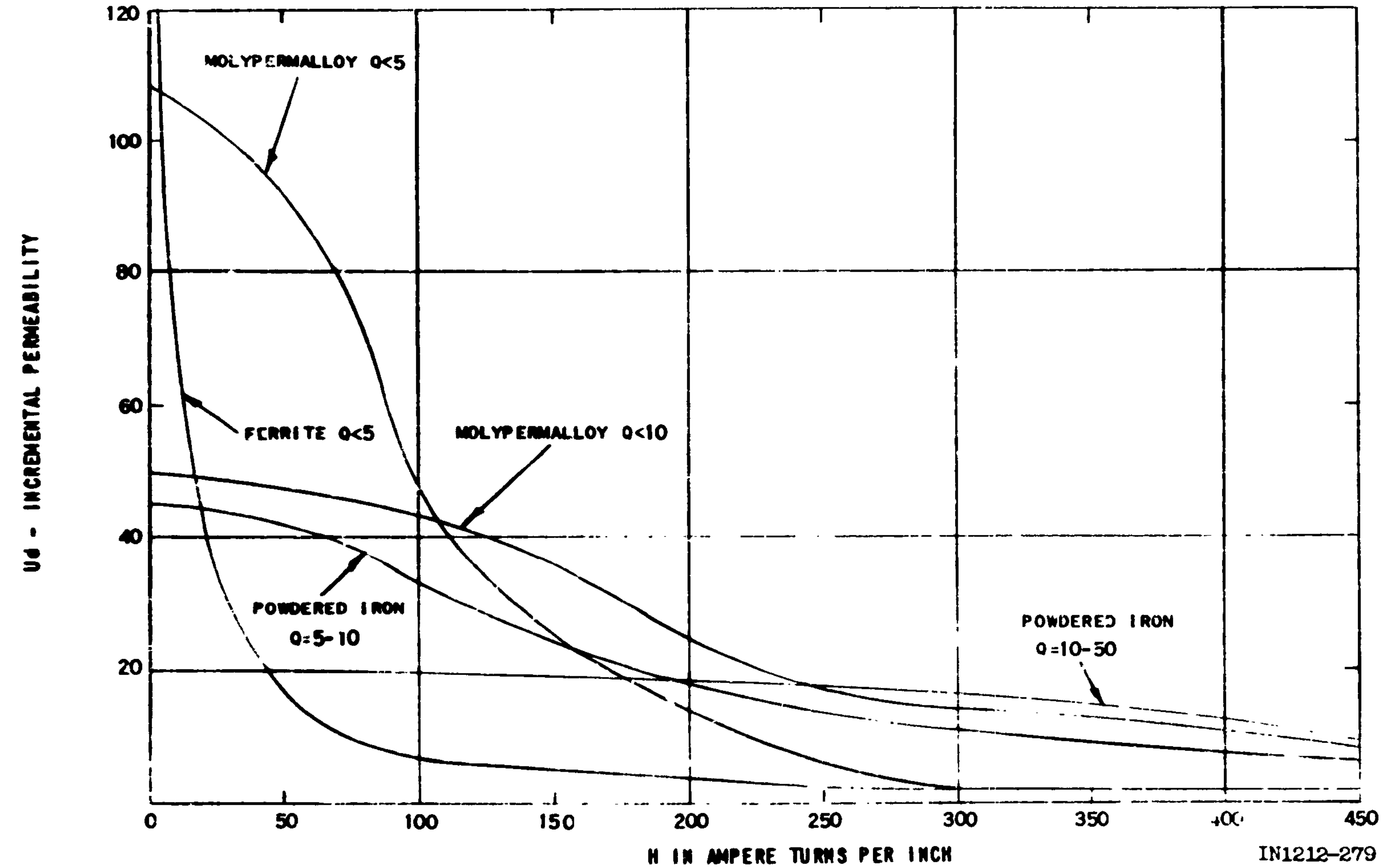


Figure 3-90. Incremental Permeability as a Function of Magnetic Field Intensity (200 kc for each Q)

power supply transformers, and isolation transformers is helpful in minimizing interference. Typical applications of these shielding techniques will aid the engineer in the design of suitable units.

b. Methods of Shielding. The interference effects upon a transformer from stray external magnetic fields are attenuated by magnetic shielded enclosures, such as Mu-metal shield-cans. As many as three Mu-metal shield-cans are used when high flux densities are encountered. The cans are nested together, and a heavy copper interleaf is used between them, resulting in attenuation of the order of 100 db. An additional 45 db of attenuation is obtainable when the transformer windings are wound in a hum-bucking configuration. These values of attenuation are dependent upon the transformer's orientation with respect to the magnetic field. The Faraday shield is employed whenever a high level of isolation between windings is required. The isolation is accomplished by enclosing the windings in one or more Faraday shields. The Faraday shield consists of a grounded area of metal that is placed between the source of the field and the winding to be shielded. The ground maintains the shield at a constant potential, thus preventing the field lines from reaching the winding. The impedance of the shield itself and its ground-lead should be as low as possible to ensure that the entire shield becomes a true equipotential surface. If the shield is to be used where magnetic field shielding and eddy-current losses must be avoided (as in shielding between the primary and secondary of a transformer), it should either be placed parallel to the magnetic lines of force and provide no shorted-turn effect, or be subdivided to break up the eddy-current paths. A subdivided shield is made by weaving a cloth of wires in one direction and threads in the other; all the wires are then connected together and grounded at one end. These shields are commercially available. Their use prevents the coupling of residual currents from primary to secondary and also provides a high impedance from the secondary load back to the line.

c. Leakage Capacitance.

(1) The high level of isolation obtained with a Faraday shield is provided by the equivalent coupling capacitance, which may be less than 0.03 pf between windings. This low-leakage coupling value gives rise to a common-mode signal rejection in excess of 130 db when the undesired interference is at a frequency less than 400 cps. The equivalent capacitance is determined by the equation:

$$C = \frac{E_2/R}{2\pi f E_1} \quad \text{(farads)} \qquad (3\text{-}47)$$

where: C = equivalent capacitance in farads

R = resistance in ohms

f = frequency (cps)

E_1 = input voltage (volts)

E_2 = output voltage (volts)

A test circuit for measuring these values is shown on figure 3-91. The achievement of a low-coupling capacitance value between windings depends upon one important construction feature: shielded leads must run directly into the transformer windings. Terminal pins are not used because the capacitances between a terminal pin and a transformer case alone can be as high as [illegible]. On installation, shields should be grounded with a low-impedance bond. If the shields are grounded to the frame of the transformer, the contact surface of the transformer base and chassis must be electrically clean.

(2) On figure 3-92 a standard transformer with one Faraday shield is shown; C_1 and C_2 represent the leakage capacitance. In this case, the equivalent capacitance can be as high as 100 pf at 60 cps, representing an impedance

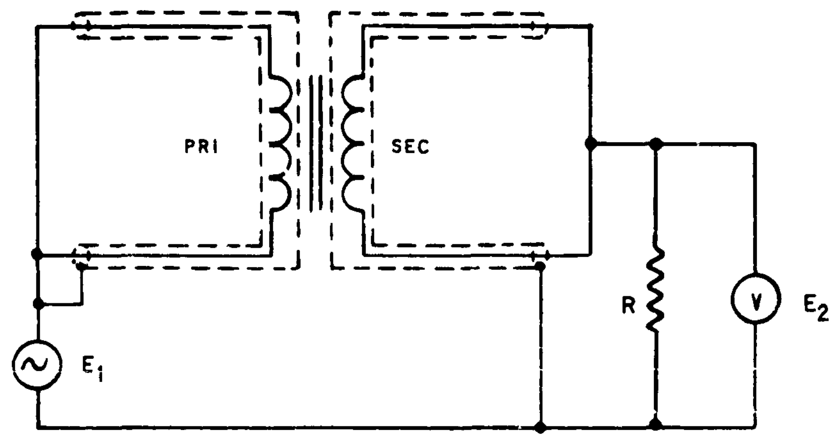


Figure 3-91. Test Circuit for Measuring Leakage Between Primary and Secondary in a Box-Shielded Transformer

of 26.6 megohms. Thus, a current of 4.33 μa will flow through the one-megohm load, producing a noise voltage of 4.33 volts. By comparison, a Faraday box-shielded transformer and its equivalent leakage capacitance of only 0.03 pf produce a leakage impedance of 8.84×10^{10} ohms. Only 1.3 mv of noise voltage appears across the load. A double Faraday box-shield around the secondary of a power transformer is often necessary when a floating dc supply is desired. As shown on figure 3-93, the isolation impedance from the output of the rectifier bridge to ground is represented by R. The capacitance (C3 and C4) from the winding to the shield shunts this value of R and reduces the isolation impedance. The use of a guard shield (fig. 3-94) makes the shunting effect negligible. Connecting the guard shield to the center-tap of the winding assures that there will be no significant potential difference between the guard shield and the outer shield. The shunting impedance is essentially infinite: current does not flow between the guard shield and the outer shield, and the capacitance is confined to the secondary winding.

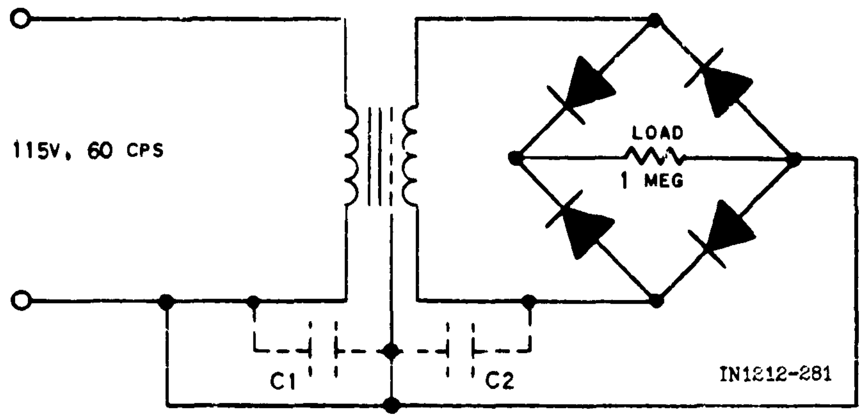


Figure 3-92. Standard Transformer with One Faraday Shield

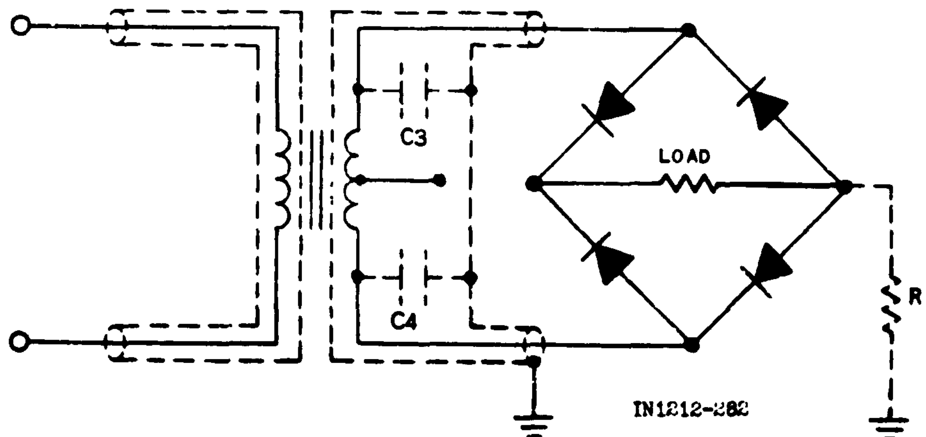


Figure 3-93. Double Faraday Box-Shielded Power Transformer

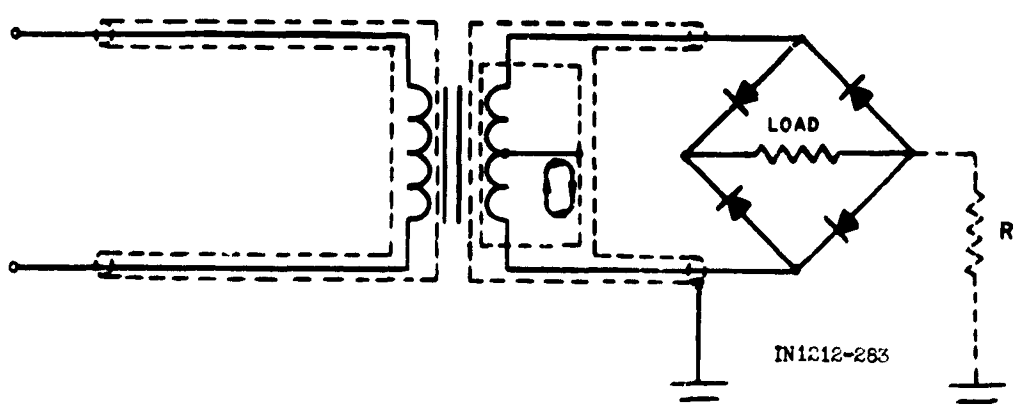


Figure 3-94. Guard Shield on Power Transformer

d. Low-Level Input Transformers. In low-level systems, the amplitudes of common-mode interference signals often exceed the signal voltages being measured. Common-mode rejection, therefore, becomes vital. The Faraday box-shielded transformer, when properly connected, offers the greatest amount of common-mode rejection. A guard shield can also be used in conjunction with an input transformer. The guard shield is installed in the primary side, rather than in the secondary as in the case of the power transformer. Usually, the guard shield is connected to the winding center tap, while the outer shield is returned to the low side of the source. Various schemes have been employed by electronic design engineers for terminating these shields. One method commonly used is that of driving the shields at a voltage potential above ground.

e. Chopper Input Transformers. The requirements of chopper input transformers are similar to those for low-level input transformers; complete isolation between windings, a high order of common-mode rejection, and rejection of stray magnetic fields.

f. Isolation Transformers. Faraday box-shielded isolation transformers have valuable application on power lines where motors, relays, and other devices are in operation, and are producing interference signals common to each line. These signals can usually be controlled, or greatly diminished, by the use of these isolation transformers. When it becomes necessary to isolate several instruments from each other within different parts of a circuit, a separate isolation transformar can be used for each instrument.

Section III. POWER SUPPLIES

3-15. General

Power supplies are essential components of most electronic equipment. Interference from power supplies can occur in two ways (fig. 3-95): Interference may be generated within the power supply and transferred to other circuits, or interference may be generated in one circuit or piece of equipment and transferred through the power supply to other circuits or pieces of equipment. The ideal power supply does not generate interference or serve as an interference transfer medium for other circuits. To achieve such a power supply, completely interference-free circuits would have to be used, and every input and output line would have to be decoupled and shielded from all external electric and magnetic fields. When it is impractical to design such an ideal power supply, other interference control measures should be used. These include filtering, shielding, circuit planning, and selection of components. The guidelines presented in this section should be adapted to the particular power supply being designed. If it is impossible to use separate power supplies, well regulated supplies are useful. The regulation causes very low common impedance coupling between circuits, which will substantially reduce this mode of interference transfer.

3-16. Circuit Planning

a. When a common power supply is used for several circuits, it presents a special compatibility problem: circuits using a common power supply must be compatible with each other. For example, if a particular piece of equipment uses high-power relays in association with low-power digital-information circuits, an interference-reduction problem exists if only a single power supply is used. The relay circuitry, because of its inherent high-interference level and sudden voltage-current demands, will completely immobilize the low-power digital circuits with their stable voltage-current requirements. Interference generated in the relay circuits can very easily be transferred through the power supply into the digital circuitry. One solution is to design separate power supplies for both the high-power relays and the sensititve digital circuitry. In this way, both circuits can be completely isolated, and the mutual coupling problem eliminated.

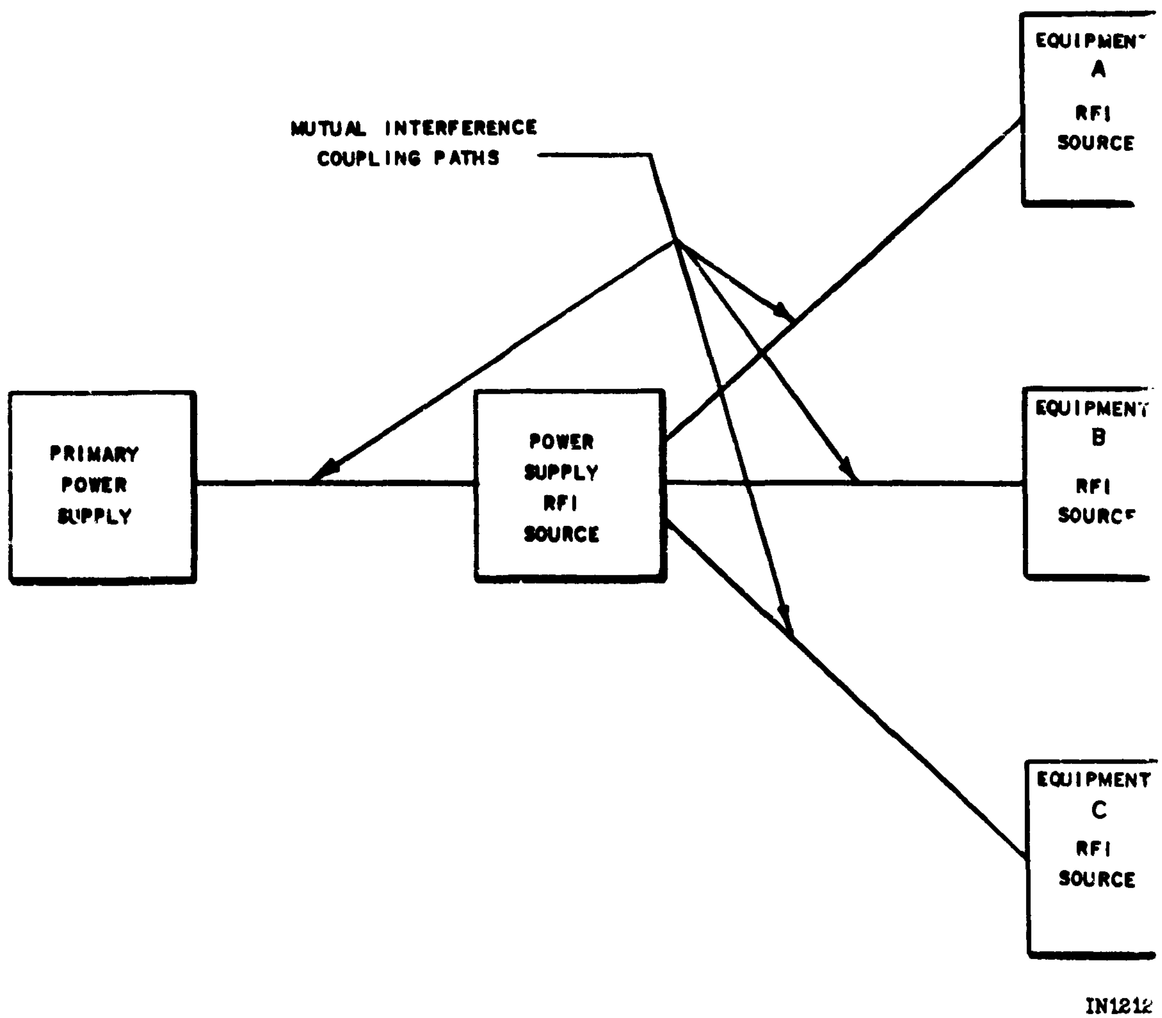


Figure 3-95. Mutual Interference Coupling

b. When a common power supply must serve many circuits, certain interference suppression techniques should be incorporated. The power supply should isolate each of the circuits, and interference sources located in each circuit should be decoupled and isolated from other circuits. The circuits should be decoupled and isolated in accordance with their individual power requirements, electrical characteristics, frequency and spectrum characteristics, and function. Circuits that produce high-interference levels should be isolated from sensitive circuits. Conducted interference, originating within one piece of equipment, should be prevented from being transferred through the power supply into common power lines and into other pieces of equipment. As shown on figure 3-95, a power supply can provide mutual interference coupling paths between all the equipment that it serves. Interference of this type can be controlled by the proper use of interference filters in the offending lines (fig. 3-96). When the power supply is an interference source (fig. 3-96A), its lines should be filtered; this prevents the interference from coupling to other equipment; instead, it remains within the power supply circuitry. When the equipment is the interference source, the equipment lines should be filtered (fig. 3-96B); the filtering isolates the interference within the equipment and prevents coupling to the power supply or other pieces of equipment.

c. The physical location of electronic circuits is an important design consideration. Theoretically, all circuits employing a common power supply should be located as close together as practicable. This arrangement often minimizes or eliminates interference problems before they arise: interference can be contained within a small area, the number of suppression components and amount of shielding can be reduced, and long lengths of shielded cable can be eliminated. Figure 3-97A illustrates a power supply which is isolated from its load. The interference control measures consist of two shielded cases, filter, shielded cables, and an internal case partition. When the power supply is located close to its load (fig. 3-97B), the only interference control measure required is one shielded case.

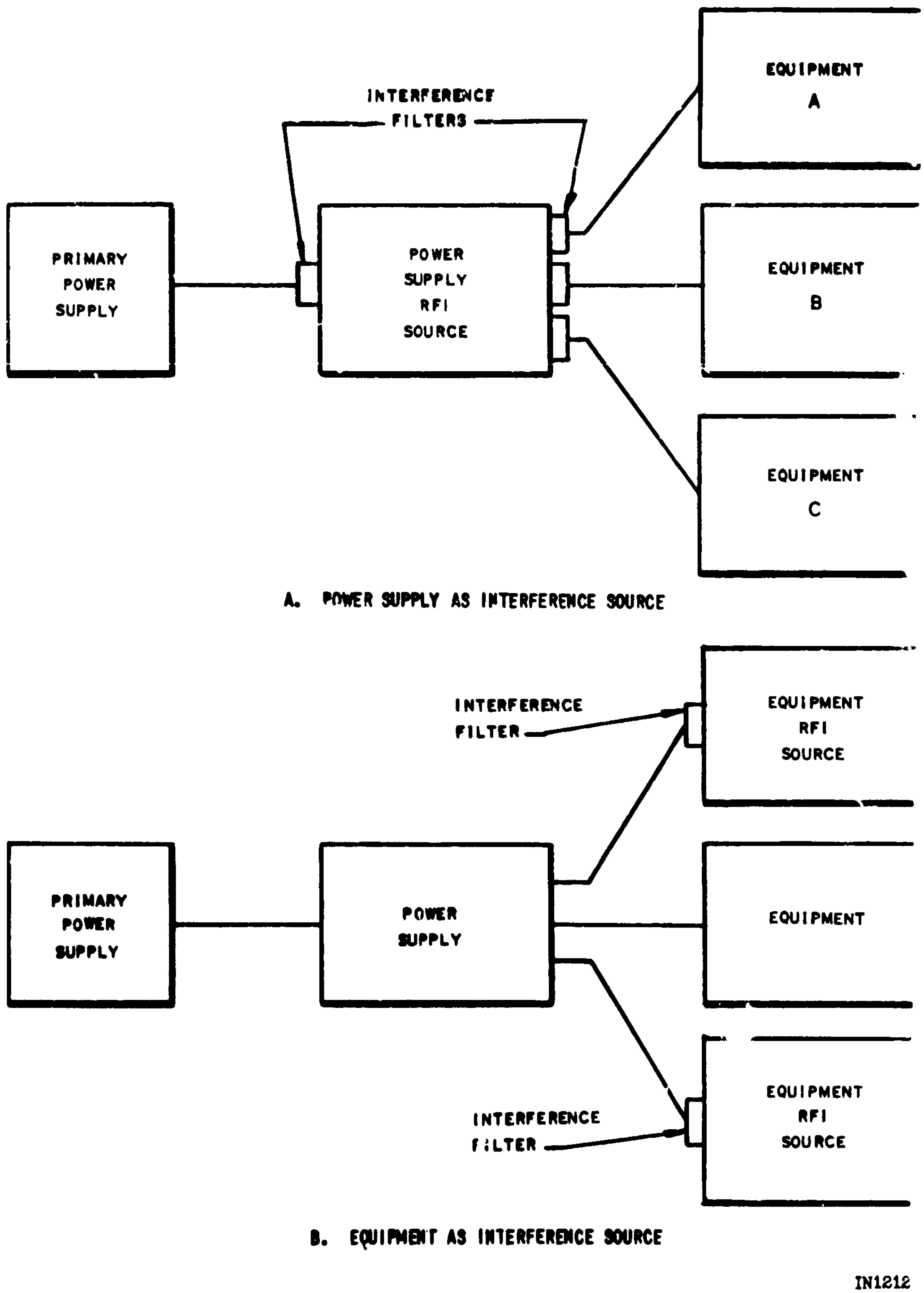
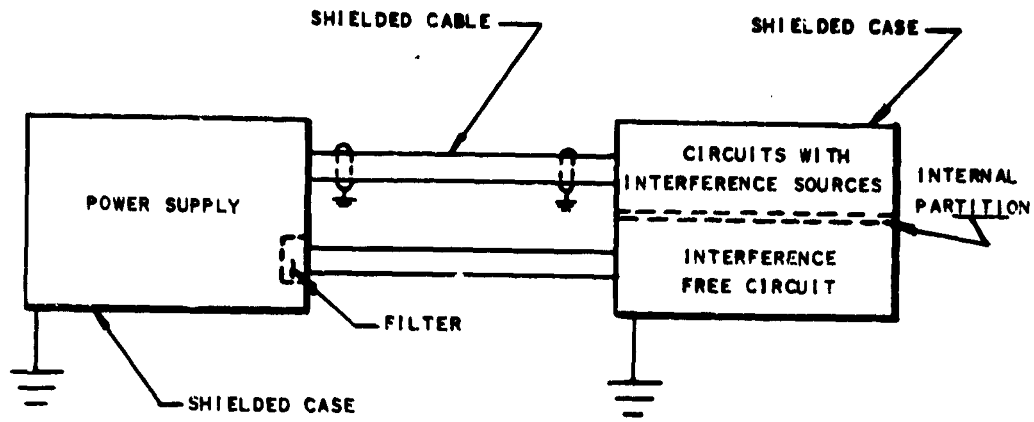
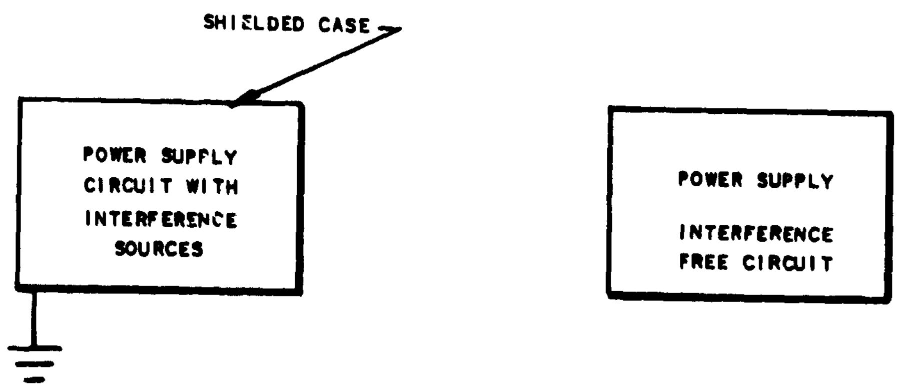


Figure 3-96. Power Supply Filtering



A. POWER SUPPLY ISOLATED FROM LOAD



B. POWER SUPPLY CLOSE TO LOAD

IN1212-11

Figure 3-97. Power Supply Isolation

3-17. Supply Circuitry

a. The components normally used in power supply circuitry, such as rectifiers, diodes, and thyratrons, are prolific sources of interference. Rectifier circuits generate broadband and power-frequency interference; the broadband interference is generated during the switching process of the rectifiers, and power frequency interference is generated by the associated transformers and choke coils. Section 1 of this chapter gives a detailed description of rectifier interference. Gas-tube and rf circu usually produce high levels of interference in power supplies. When gas diodes and thyratron rectifiers are used, interference arises from two distinct effects: the steep voltage and current wave-fronts asso iated with the firing (ionization) cycle of the tube, and the plasma oscillat or during the discharge. The external effects of both types of interferenc can be minimized by shielding the tube or the complete circuit and by rf filtering. The rf filtering is used in addition to the heavy low-frequency filtering of all leads.

b. Radio-frequency circuits comprise free-running oscillators and sharply resonant circuits triggered by pulses. In both cases, the oscillator output is amplified and then rectified and filtered. The interference consists of rf energy which radiates out from the tube, transformer, and other components, and is conducted along connecting wir Where it is impossible to avoid the use of rf high-voltage rectifiers, t interference must be confined to the power supply itself by proper and adequate shielding and filtering.

Section IV. CONTROL CIRCUITRY

3-18. General

Control circuitry that utilizes electrical, electronic, or mechanical switching devices can produce considerable amounts of broadband interference throughout the electromagnetic spectrum. Such interference is generated during the transient (electrical) state immediately following an abrupt change in a control device. Although the various types of circuits described here all reduce generated interference, there are no specific design procedures that can be followed which will guarantee complete interference suppression. There are an unlimited number of interference reduction circuits that can be used. For complicated interference reduction problems, it is often more expedient to apply filter networks instead of experimenting with interference reduction circuits.

3-19. Over-All Interference Reduction in Switch Circuitry

a. Under normal operating conditions, the establishment or interruption of current flow through a switch cannot be achieved in a smooth transition between the two steady states. Upon switch closure, final establishment of firm contact is preceded by an interval of premature electrical closures, or bridging, between the contacts. At the very first instant of current flow, the circuit inductance resists any change in current, and extremely high voltage is built up across the switch gap at a very rapid rate. This voltage creates an electric field intense enough to melt the contact surfaces of the switch electrodes and draw a molten metallic bridge across the gap. This same high-field condition arises during the initial phases of switch opening in the interval when the contacts are in very close proximity. In this case, numerous closures result before final interruption of the current is achieved. In addition, when a switch is opened in an inductive circuit, the transient voltage appearing across the switch gap may be sufficient to cause arcing and a gaseous discharge, referred to as glow discharge, or sawteeth.

b. Switch-generated interference is related to three causes: high-voltage gaseous discharges (sawteeth); low-voltage, high-field breakdowns (bridging); and current changes between the two steady-state values. Th high-voltage sawteeth, when present, produce more interference than all other causes combined throughout the entire frequency range from 15 kc to 1000 mc. There is a broad peak in the spectral distribution of the interference caused by these sawteeth that occurs in the region of a few megacycles for common values of circuit parameters. The bridging has a negligible effect at the lowest frequencies, but increases in relative importance with frequency until, at 1000 mc, it is nearly as great a source of interference as the sawteeth. The change of current between the two steady-state values yields its greatest contribution to interference at the low-frequency end of the spectrum (fig. 3-98).

c. In reducing interference, first consideration should be given to elimination of high-voltage sawteeth because of the great amplitude and wide spectral distribution of the interference associated with these discharges. This is accomplished by employing an interference reductio device that prevents the voltage across the switch gap from exceeding th value required to initiate a glow discharge (approximately 300 volts). To eliminate the formation of bridges, it is necessary to prevent the electric field between the contacts from exceeding a critical value (approximately 5 x 10^6 volts-per-inch for untreated contacts). Bridge elimination may be accomplished upon switch opening by mechanically increasing the speed of separation of the contacts, and by electrically decreasing the rate of build-up of the potential across the contact gap. It is impossible to alter the changing load current between the two steady-state values without affecting normal operation of the circuit; however, interference may be effectively contained in the regions of the switch and load side of the circuit by inserting filters in the externa power supply leads. A good interference reduction circuit should:

1) Retard the build-up of voltage across the gap during the init period of contact separation to minimize bridging reclosures

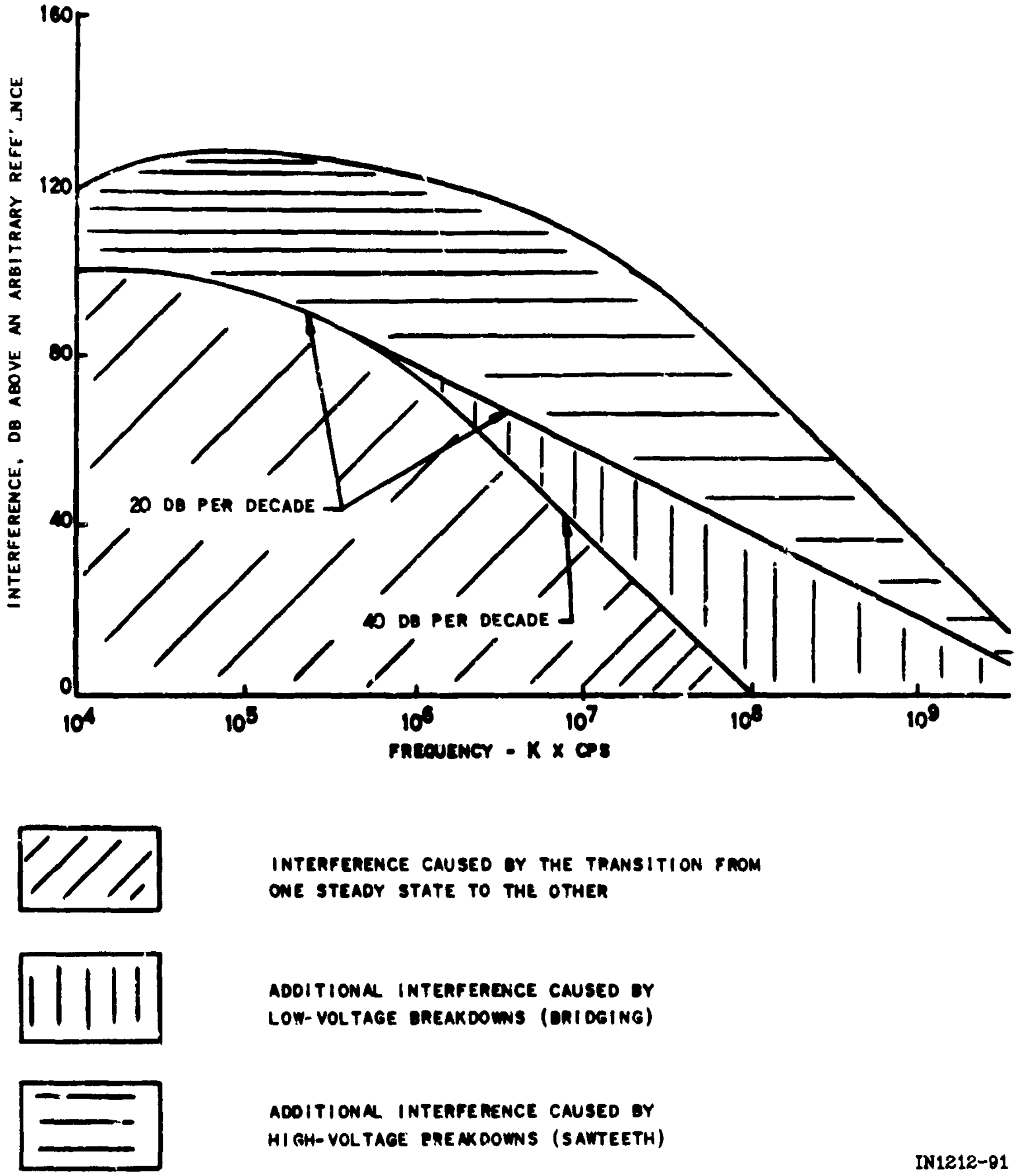


Figure 3-98. Summary of Switching Interference Phenomena

2) Limit the peak voltage across the switch gap, upon opening, to eliminate gaseous discharges

3) Limit the surge of current through the switch, upon opening or closing, to minimize sharp wave-front transients

d. In designing an interference reduction circuit, two steps should be followed: selection of the circuit and determination of the values of the components in the selected circuit. In circuit selection, it is usually necessary to make a trade-off between interference reduction and other considerations, such as the allowable decay time of the load curren ease of installation of the interference reduction circuit, physical size of the interference reduction components, and adaptability to the power supply. The circuit that gives the greatest interference reduction, a resistor in series with the switch and a capacitor in shunt with the swit resistor unit, will extend the decay time of the load current upon switch opening. The capacitor often becomes physically large when the energy stored in the inductive field of the load is great. The rectifier-bias-battery circuit, which has a very small decay time, gives poor interfere reduction because it cannot reduce bridging. The load-shunt diode circui as shown on figure 3-99, changes the circuit decay time.

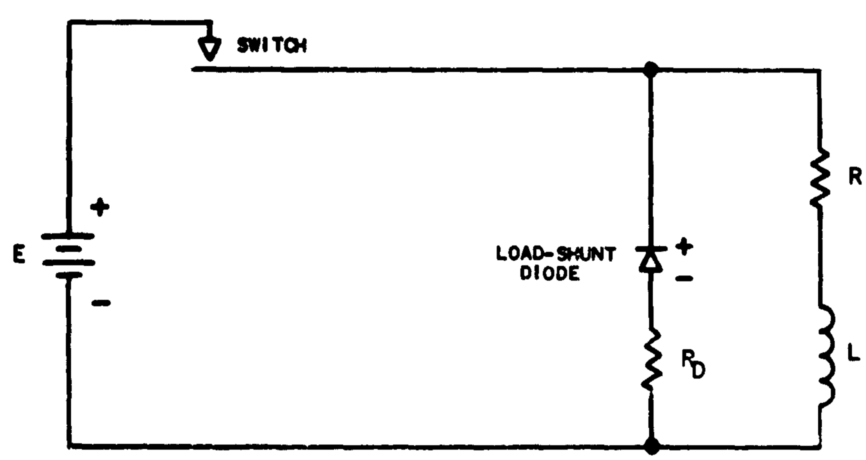


Figure 3-99. DC Switching Interference Reduction By Load-Shunt Diode

e. After selection of the interference-reduction circuit to be used, component values must be selected, based on a knowledge of the supply voltage, load current, load inductance, and frequency of switching. For example, for a capacitor placed across the switch, the value of the capacitance is determined by the current value and the load inductance -- supply voltage and frequency of switching not being important factors. It is necessary that such a capacitor be able to store temporarily all of the energy previously contained in the load inductance without the voltage across the capacitor exceeding the glow discharge value of approximately 300 volts $\left(C>L(I/300)^2\right)$. In the load-shunt diode circuit of figure 3-99, the voltage at which the knee in the reverse characteristic occurs must exceed that of the supply. The diode need dissipate only a fraction of the total energy stored in the load, as most of it is dissipated within the load resistance itself. For rectifier applications, a conservative selection of the diode requires an average current rating equal to the load current to be interrupted. In applications of low-duty cycle, the diode may be up-rated by a sizeable factor. In ac applications, when back-to-back diodes are used, most of the stored energy is dissipated in the reverse diode. This diode, therefore, must be selected on the basis of heating. If the duty cycle is high, it may well be that this diode will have (for this application) a current rating that is less than its normal rectifier rating.

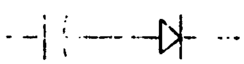
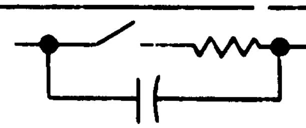
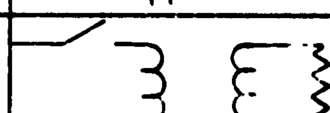
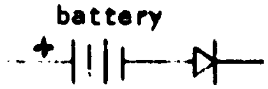
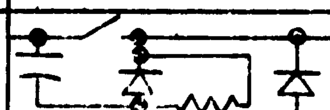
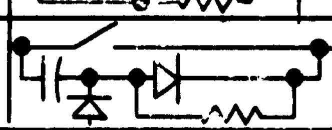
3-20. Interference Reduction Component Characteristics

a. Inductors. There is very little advantage in employing a single inductor, in series or in parallel, as an interference suppressor. Placing an additional inductor in series with a circuit is undesirable because it serves only to increase the inductively stored energy that must be accommodated upon opening of the switch. Shunting of any of the three basic circuit elements by an inductor is also useless in reducing the severity of the disturbance that takes place within the switch gap upon opening of the circuit. Placement of the inductor across the switch prevents normal operation of the circuit because it becomes impossible to reduce the load current to zero. Placement of the inductor in parallel with the load serves only to increase the inductively stored energy and, consequently, the current which the switch must interrupt. Placement of the coil in parallel with the source causes excessive current to be drawn from the supply (depending on the re· sistance of the inductor) and does not affect the nature of the phenomena that occur within the gap. An inductor can, however, be used in conjunct with other components in forming an interference-reduction circuit.

b. Capacitors. Retardation of gap voltage build-up upon opening of switch may be achieved by connecting a capacitor across the switch or ac the load. By choosing a sufficiently large capacitance, the rate of swit voltage build-up may be reduced to any desired value. Such a capacitor however, may present serious difficulties with regard to the other requirements (table 3-5). Limitation of peak voltage to a value that will not cause gaseous discharge requires that the capacitance be sufficiently large to accommodate all of the energy which, prior to the switch openi was stored in the inductive field of the load. With loads having large inductances and large operating currents, the required value of capacita may be far too large for practical application. The third requirement of table 3-5 is not satisfied with either positioning of the capacitor. C of the switch results in a very large discharge or charging current, wit the consequence that contact erosion is very rapid. In addition, if the capacitor is placed across the load, this very large peak of current must flow through the supply wires and may be a source of interference to adja circuits. With the capacitor placed across the switch, this surge of cur is confined to the small loop consisting of switch and capacitor only. T interference from the supply wires is reduced, but contact erosion is sev

c. Resistors. Some reduction of interference may be accomplished by placing a linear resistance across either the load or the switch. In pra there is a lower limit to the value that this resistance may have for bot these positions. With the resistor shunted across the switch, a lower li is placed on the load current since opening the switch will no longer red the current to zero. On the other hand, this resistor, when placed acros load, becomes an additional wasteful load that the source must supply whe ever the switch is closed. The gap voltage is also the peak value of the

TABLE 3-5. COMPARISON OF INTERFERENCE REDUCTION COMPONENTS

Components	Placement	Requirements: 1 Retard Build-up of Gap Voltage	2 Limit Peak of Gap Voltage	3 Minimize Sharp Wave Front Transients
Capacitor	Load	G	A[c]	P
	Switch	G	A[c]	P
Linear resistor	Load	P	A[c]	A
	Switch	P	A[d]	G
Semiconductor diode	Load	P	G	A[b]
	Switch[a]	P	G	
Back-to-Back diodes	Load	P	G	A[b]
	Switch[a]	P	G	A[b]
Capacitor and diode 	Load	Capacitor is superfluous		
	Switch	G	A[c]	G
Series R shunt C		A[d]	A[c]	G
Coupled secondary		P	G[d]	A
Diode and battery 	Load	P	G	A[b]
	Switch	P	G	G
Composite Circuit		G	G	A
Composite Circuit		G	G	G

G = Good A = Intermediate P = Poor

a = Diode must have knee at voltage greater than that of supply

b = Determined by inherent shunt capacitance of diode

c = Capacitance must be sufficiently large

d = Resistance must be sufficiently small potential that appears across the opening switch. To avoid gaseous discharges, this voltage should not exceed a value of 300 volts. If placed across the switch, the gap voltage is equal to the product of the resistance and the interrupted current. When placed across the load, the gap voltage is equal to the sum of the supply voltage and the product of the interrupted current and the suppressor resistance. Thus, an upper limit is imposed on the allowable range of values for the resistance. Placing the resistor across the switch will have less of an adverse effect because no surge of current can occur upon closure of the switch. Instead, the current through the source, switch, and load builds up exponentially because of the load inductance. With the resistor across the load, the source and switch current immediately assume the value drawn by this resistor from the source. While this current may be large and may begin to flow abruptly upon switch closure, it is less severe than the large impulse of current that will flow if a capacitor is substituted for the resistor across the load. When a resistor is placed across a switch in series with an inductive load, it is possible to generate a strong sawtooth voltage. If there is feedback anywhere in the circuit, sustained interference may result.

d. Diodes and Varistors. Devices with nonlinear resistance-voltage characteristics, such as diodes and varistors, are useful components for interference-reduction. A diode has low forward resistance and high reverse resistance. Consequently, it may be used to present either a short-circuit or an almost infinite impedance, depending upon the direction of current flow. A varistor conducts well at high voltage but not at low voltage. It is a nonlinear resistance which is very high at low voltage, but drops to a very low value at high voltage. The function of either a diode or a varistor in an interference reduction application is to provide an alternate shunt path for the induced current that presents a lower resistance than the contact gap.

(1) DC Circuits. An effective interference reduction element is the diode shown on figure 3-99. The diode is inserted in the circuit so that its polarity opposes that of the impressed

voltage. The I^2R_D loss (R_D is the resistance of the diode) through the shunt circuit is small. For maximum interference reduction, the diode should be installed as close to the offending element as possible, preferably by incorporating the diode into the relay coil or solenoid unit. Including the diode as an integral part of the load automatically builds interference reduction into the switching circuit, independent of the switching contacts used. If it becomes necessary to replace the contacts, interference reduction is not disturbed. In this position, negligible current flows through the diode under steady-state conditions, but, when the switch is opened, the diode provides a low-resistance circuit through which the inductive current may flow. The peak voltage appearing across the switch is thereby limited to the sum of the battery voltage and the forward drop of the diode. By choosing a diode with a low forward resistance, the forward drop of the battery can be made very small, so that the voltage across the switch is essentially just that of the battery. This circuit is very effective in providing the complete elimination of all high-voltage, sawtooth discharges; it is ineffective in retarding the gap voltage build-up. A second method of using the diode is shown on figure 3-100. This circuit requires that the diode possess a sharp knee in its voltage-current curve at a voltage value that is equal to, or somewhat greater than, the supply voltage so that the steady-state current through the load will be essentially zero when the switch is open. The proper location of this knee can be obtained by using a suitable Zener diode.

(2) AC Circuits. Components consisting of two similar diodes placed back-to-back are commercially available as a single unit. No matter how the unit is employed, the voltage-current characteristic of such a unit is determined primarily by the reverse diode. When placed across the load, the back-to-back unit is inferior to the load-shunt diode interference reduction circuit in that it does

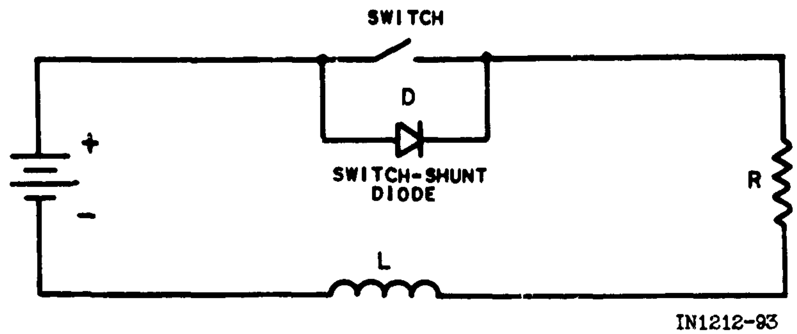


Figure 3-100. DC Switching Interference Reduction By Switch-Shunt Diode

not limit the overshoot in the gap voltage to as small a value. Bridging is therefore more prevalent. Because the unit is insensitive to the polarity of the supply, it is not necessary to check polarity when installing it in either an ac or dc circuit. Two diodes are used, as shown on figure 3-101. Each diode inhibits current flow from the source voltage when its polarity opposes that of the source. When the contacts are opened, one diode, depending on the polarity of the source at the instant switching occurs, will limit the peak driving voltage of the coil. The stored energy of the coil is dissipated in the resistance of the rectifiers and in the coil itself.

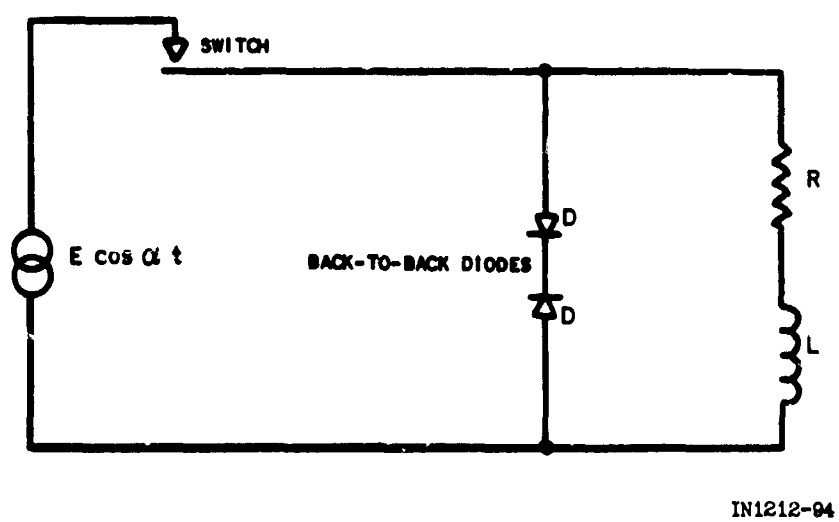


Figure 3-101. AC Switching Interference Reduction By Two Diodes

(3) Methods of Connecting Diodes and Varistors. Several arrangements of diodes and varistors for interference reduction are illustrated on figure 3-102. Circuit A shows the diode across the load in such a way that, when the relay contact is closed, it conducts and causes high battery drain. It has high resistance when the relay contact is open. This location is not a good one for a diode. A varistor would be more advantageous because it would conduct at high induced voltage with relatively low battery drain at normal circuit voltages. Circuit B results in low power consumption and minimum heating. The low resistance is in the direction of surge current so that it is a better arrangement than A for interference reduction. A diode is more effective than a varistor in this position because it conducts at a lower induced voltage. The battery drain under steady-state conditions is the same as with circuit A when using a varistor. Circuit C presents a high resistance to both the supply and induced voltages. Some power is lost at all times when the circuit is open; therefore, this is not a satisfactory position arrangement. Circuit D causes high power consumption, and the circuit is never completely open. A diode, capable of dissipating the power under repeated operation, allows about 20 percent of normal current to flow at all times. A varistor can be used here; it would reduce the steady-state power drain. Circuit E minimizes delay in voltage decay for special purposes and can be used in both ac and dc circuits because it blocks voltages in both directions. It is required that the knee in the diode curve be above the normal circuit voltage. This arrangement provides low power loss when the circuit is open; there is an increase in power loss when it is closed. Circuit F is very useful

because it dissipates little power under steady-state conditions, and none when the circuit is open. It is very effective in interference reduction when the same design factors specified for circuit E are employed.

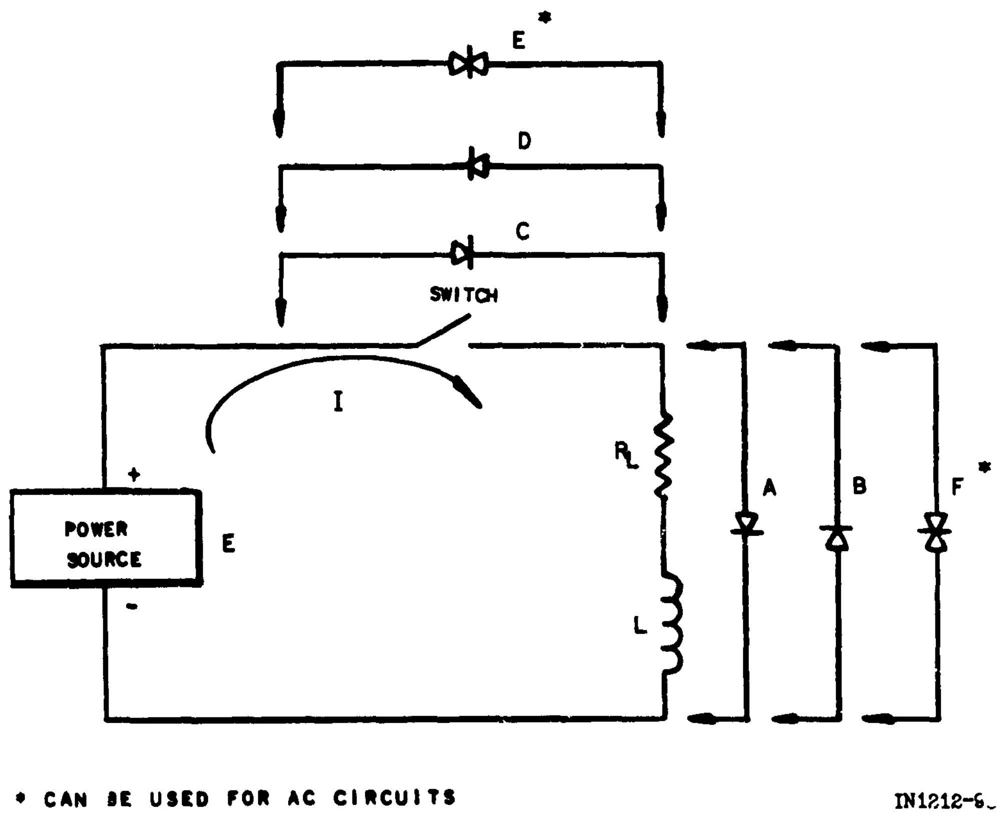


Figure 3-102. Methods of Connecting Diodes and Varistors

(4) Application Guidelines. When designing interference reduction circuits using diodes or varistors, the following design guidelines shouod be considered:

1) When using a single diode, the rectifier current-rating should be equal to the load current when repeated cycling is necessary

2) For intermittent use, a rating equal to one-half of the load current is usually sufficient

3) Peak inverse-voltage ratings of silicon and germanium diodes should exceed the supply voltage by at least 20 percent. For selenium diodes, a lower safety factor is normally satisfactory

4) Back-to-back diodes may be used for ac circuits. The peak inverse-voltage rating must exceed the supply voltage

5) When using varistors, the characteristic curve resistance knee must be above the supply voltage, and the heat dissipating area must be designed sufficiently large

3-21. Interference Reduction Circuits

a. Series Capacitor and Resistor. Contact erosion, arising from the use of capacitors, can be alleviated by the addition of a series resistor (fig. 3-103). The value of this resistance must be one of compromise. Because the voltage that appears across the switch immediately upon opening is equal to the product of this resistance and the interrupted current, this resistance should be low. On the other hand, a large value of resistance is desirable to minimize the contact erosion upon closing. While the IR drop does allow some bridging to occur, a sufficiently large value of capacitance will eliminate all high-voltage sawteeth. The values of resistance and capacitance should be empirically chosen from the results of test data to give a minimum of interference.

b. Series Capacitor and Nonlinear Resistance. The series capacitor and resistor can be improved by shunting the resistor with a diode, as shown on figure 3-104. This action permits the achievement of the desired low-resistance value for a switch opening and the desired high-resistance value for switch closing. For maximum effectiveness, it is necessary that the capacitor completely discharge during the closed interval of the switch, or bridging will occur when the switch is again opened. It is therefore necessary to maintain the linear resistance in shunt with the diode to assure that the initial condition is maintained.

c. Typical Resistance-Capacitance Circuit. A very effective circuit, shown on figure 3-105, consists of a resistor placed in series with the load circuit and a capacitor in parallel with the series combination of resistor and switch. This circuit is similar to the series capacitor-resistor combination. Because the resistor must carry the normal load current with a negligible voltage drop, there is a practical limit to the maximum value of the resistance. (This requirement is in addition to those imposed by the interference reduction considerations.) In practice, this circuit is much more effective in reducing noise than the series capacitor-resistor unit. Whatever disturbance is produced is largely confined to the switch-resistor-capacitor loop. This circuit not only alters the phenomena occurring at the gap, thereby functioning as an interference reducer, but also provides containment for the interference which is produced.

d. Special Interference Reduction Circuits. Special interference reduction circuits are those that deviate from the basic circuit of switch, supply, and load, all connected in series.

(1) Parallel-switch circuits. A parallel-switch circuit that employs an inductive load but exhibits an interference spectrum characteristic of a resistance load is shown on figure 3-106. Considering the circuit in a steady state with the switch initally in the open position, the current (I_L) through the load inductance (L) and the source current (I_S) are identical. The voltage drop (V_S) across the switch is one-half of the source voltage (V). When the

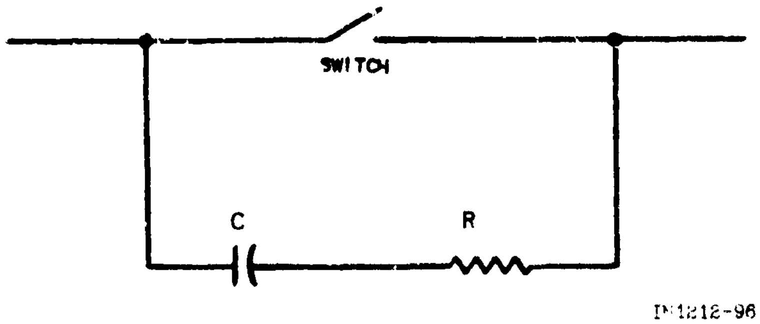


Figure 3-103. Series Capacitor and Resistor Circuit to Reduce Contact Erosion

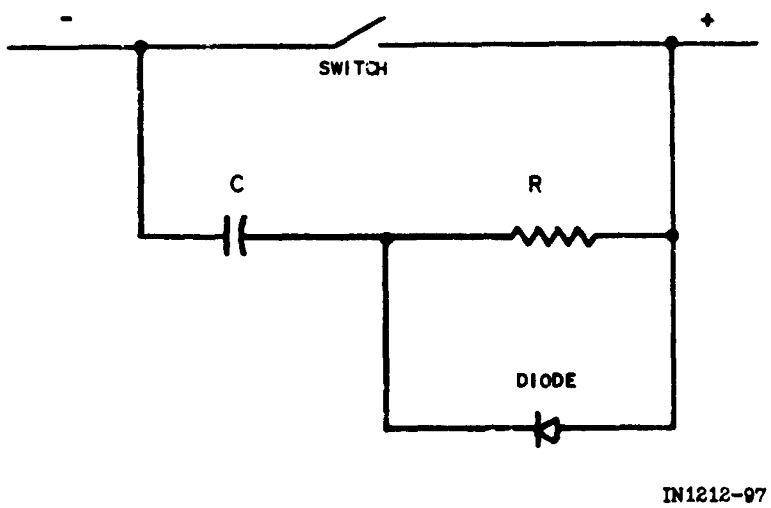


Figure 3-104. Interference Reduction by Series Capacitor and Non-Linear Resistance

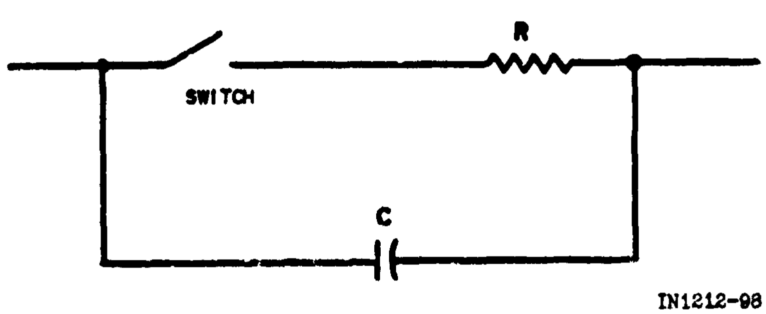


Figure 3-105. Interference Reduction by Series Resistance and Parallel Capacitance

switch closes (assuming that no bridging or bouncing of the contacts occur), the switch voltage (V_S) drops immediately to zero; the source current (I_S) abruptly rises to a new value equal to twice the original, as a step function; and the inductance current (I_L) decreases, in an exponential manner, to zero. If the existence of bridging is ignored, the opening of the contacts results in the source current (which is initially zero because of the load inductance) immediately becoming equal to the load current. This current then exponentially approaches the constant operating value. The accompanying switch voltage immediately assumes a value equal to the full supply voltage upon opening of the switch, and then decays exponentially to a value of one-half that of the supply. This cycle of events is depicted graphically on figure 3-107. The parallel-switch circuit has essentially the same interference-producing mechanism as the more conventional series-switch circuit with a purely resistive load. In both cases, step functions of switch voltage and circuit currents exist, and there is the possibility of bridging and arcing. In neither case, however, can high-voltage gaseous discharges occur.

(2) Coupled Coils. In a series circuit consisting of a dc source a switch and an inductive load, violent circuit readjustments, that result in interference, occur during the opening of the switch. The energy stored in the magnetic field of the inductance tends to maintain the flow of current through the opening switch. This readjustment may be facilitated by providing an alternate closed path through which the collapsing magnetic field can cause the current to circulate (fig. 3-108). In such a circuit, the optimum value for the coefficient of coupling, k, is unity. If k has any other value than one, there will be a primary leakage inductance that will tend to maintain the flow of current in the primary loop when the switch is opened. It is this current which should be eliminated. A transient analysis of this circuit,

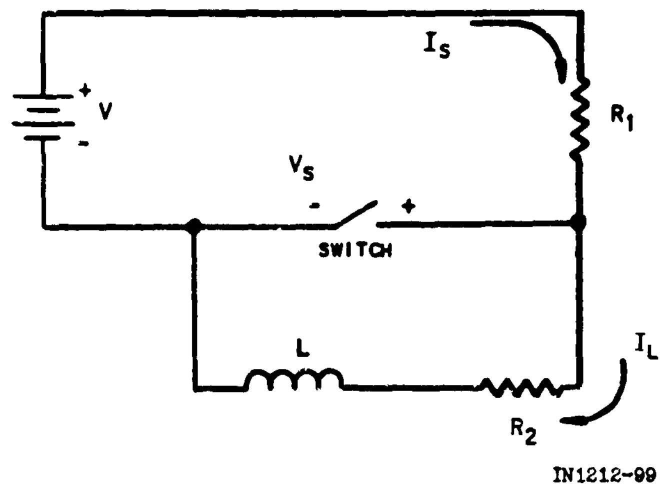


Figure 3-106. Parallel-Switch Circuit

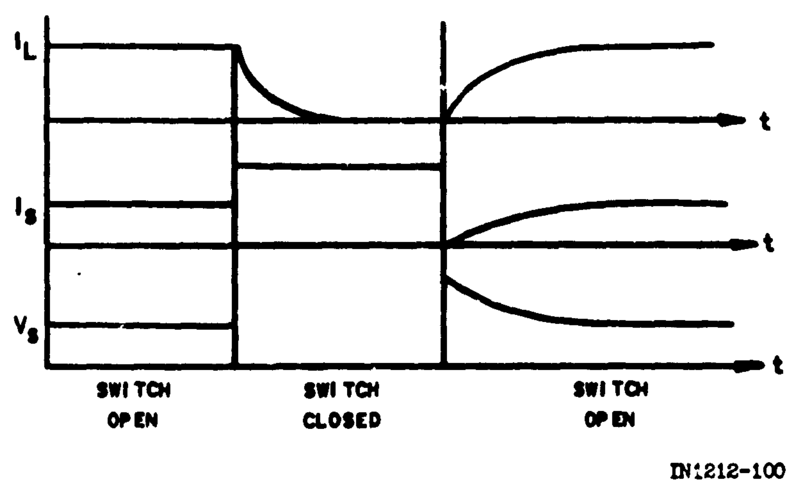
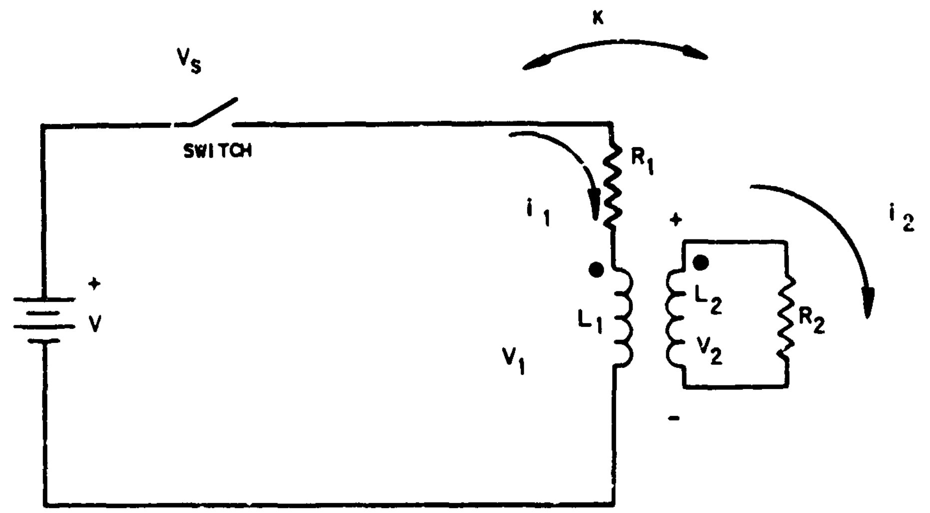


Figure 3-107. Waveforms for Parallel-Switch Circuit



R_1 = RESISTANCE OF RELAY COIL, WHICH IS THE LOAD OF SWITCH S

L_1 = INDUCTANCE OF RELAY COIL, WHICH IS THE LOAD OF SWITCH S

R_2 = RESISTANCE OF COUPLING COIL

L_2 = INDUCTANCE OF COUPLING COIL

IN1212-101

Figure 3-108. Interference Reduction by Coupled Coils

for all values of the time following opening of the switch, shows the voltage across the primary winding (v_1) to be given by:

$$v_1\ (t) = -(RL_1/L_2)\, i_1(0^-) e^{-Rt/L_2} = -R(N_1/N_2)^2 i_1(0^-) e^{-Rt/L_2} \quad (3\text{-}48)$$

where: $i_1(0^-)$ = the value of the interrupted supply current. The corresponding expression for the current in the secondary loop is given by:

$$i_2\ (t) = (L_1/L_2)^{1/2} i_1(0^-) e^{-Rt/L_2} = (N_1/N_2)\, i_1(0^-) e^{-Rt/L_2} \quad (3\text{-}49)$$

Equation 3-48 shows that, if the factor $R(N_1/N_2)^2$ is made to approach zero (either through the choice of secondary resistance or of turns ratio), the voltage across the primary winding will approach zero; and the voltage across the switch will be that of the supply. This circuit exhibits the characteristic of a resistance load in the no high-voltage sawteeth are likely to be present. An analysis of the behavior of this circuit, when the switch is closed at t = 0, shows that the current in the primary circuit is given by the following:

$$i_1\ (t) = (V/R_1)\ (\ 1 - RL/RL_1 + R_1L_2)\ e^{-RR_1\ t/(RL_1+R_1L_2)} \quad (3\text{-}50)$$

The current i_1 (t) consists of an instantaneous rise (at t = 0) of $v/(R_1 + L_1R/L_2)$ or $V/(R_1 + R(N_1/N_2)^2$ amperes, followed by an exponential rise to a steady-state value of V/R_1 amperes. To reduce the severity of the interference accompanying the closing of the switch, the height of this rise at t = 0 should be minimized. This may be accomplished by making the term $R(N_1/N_2)^2$ as large as possible. This requirement, however, is in direct opposition to the requirements for maximum interference reduction upon opening of the switch. It should be possible, nevertheless, to select a compromise value for this term that would make the opening and closing interference levels equal and permit a degree of interference reduction. Another possible solution to this conflict of requirements for the $R(N_1/N_2)^2$ term is to replace the resistor, R, by a diode to present a low-resistance value to the opening transient current and high resistance to the closing transient current. This is possible because these two transient currents flow in opposite directions in the secondary circuit. Such an arrangement is illustrated on figure 3-109.

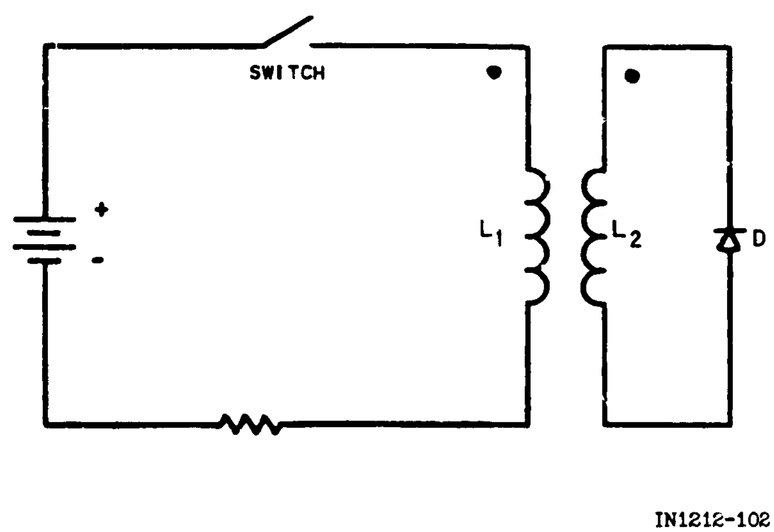


Figure 3-109. Interference Reduction by Coupled Coil with Diode

(3) Bias Batteries. In dc inductive circuits, reduction of arcing at contacts can be achieved by employing bias batteries, or their equivalent, in addition to nonlinear resistances or diodes. The primary advantage of these circuits is that the diodes need dissipate but an incidental portion of the inductively-stored energy; the diodes may therefore have lower electrical ratings and be of smaller physical size. Six variations of this type of circuit are shown on figure 3-110. Circuits A and B are essentially the same: one employs a tapped inductive load and the other a load with a closely coupled second winding. The dc source in each case also serves as the bias battery. Circuit C employs a simple load winding, but requires a separate bias battery. Circuit D is similar to C, except that the tapped winding allows greater freedom in selection of the bias voltage. Circuit E is applicable when the switching occurs at a reasonably rapid and constant rate so that the repeated transient currents through the rectifier build up a nearly constant bias voltage across the parallel resistor-capacitor combination and eliminate the need for a separate battery. Circuit F is adapted for loads that are supplied from an ac source. The basic circuit that applies for all values of time following the opening of the switch is shown on figure 3-111. In this circuit, the voltage source shown is the bias battery, and not the source that originally energized the load; I_o represents the initial value of the circuit current and is equal to the interrupted load current for circuit C. For purposes of analysis, the forward resistance of the diode is assumed to be a constant, R_R, and, when added to the load resistance, R_L, is given by R. The back resistance of the diode is assumed to be infinitely large. The circuit for the solution of L(t), for as long as the current continues to flow in the opposite direction to the reference, is shown on figure 3-112. The solution for i(t) (fig. 3-112B) can be expressed as:

$$i(t) = \frac{E}{R} - \left(\frac{E}{R} + I_o\right)e^{-(R/L)t} \text{ for } 0<t<t_1 \qquad (3\text{-}51)$$

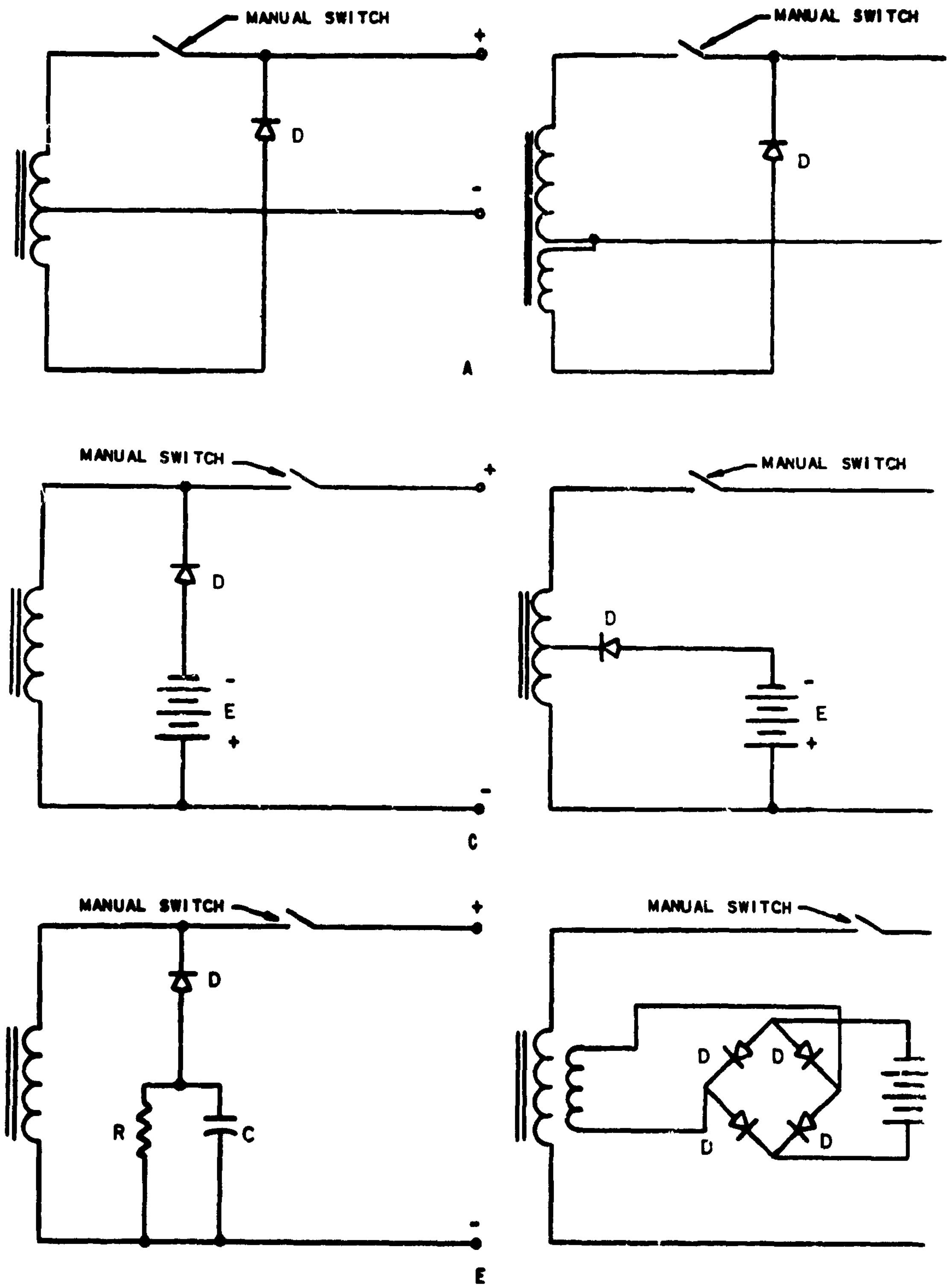


Figure 3-110. Interference Reduction Circuits Employing Bias Br

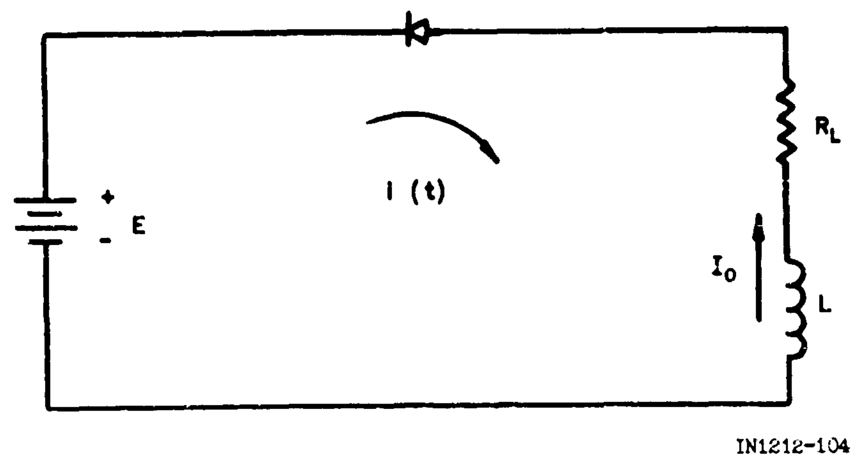


Figure 3-111. Basic Circuit Employing Bias Battery

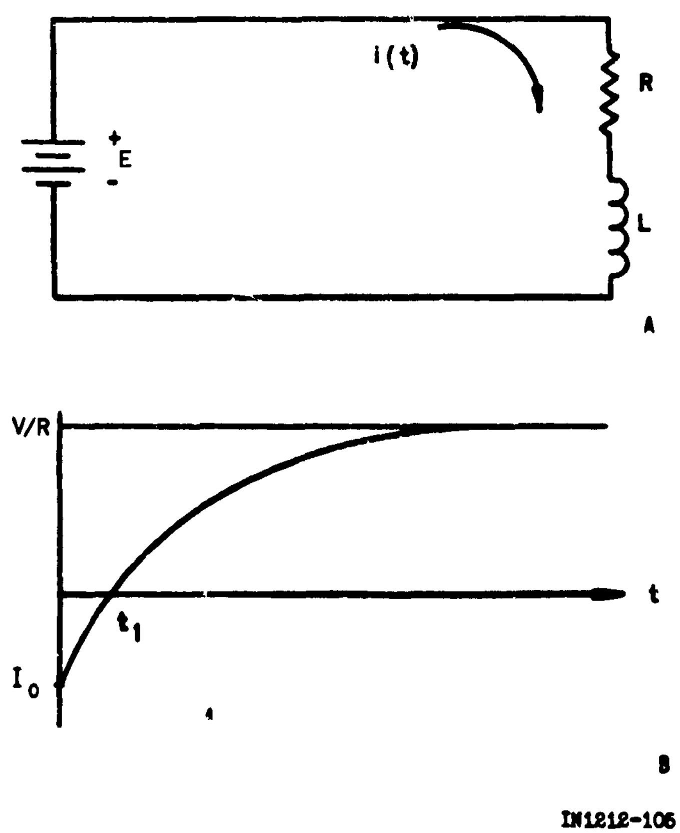


Figure 3-112. Equivalent Circuit and Plot of Current Decay

where: $i(t) = 0$ for $t_1 < t$; and t = the time at which the current passes through zero and the diode becomes an open circuit, thus preventing any further current flow. The value of t_1 may be obtained from:

$$i(t_1) = 0 = \frac{E}{R} - \left(\frac{E}{R} + I_o\right)e^{-(R/L)t_1} \qquad (3\text{-}52)$$

and shown to be:

$$t_1 = \frac{L}{R} \ln \left(1 + \frac{I_o}{E}\right) = \frac{L}{R} \ln \left(1 + \frac{I_oR}{E}\right) \qquad (3\text{-}53)$$

From this relation for t_1 (the time required for the load curren· to decrease to zero value), it can be seen that increasing the b· voltage materially decreases this time interval (fig. 3-113). · instance, if the bias voltage is equal to the product of the nor· load current value and the circuit resistance -- which is usuall, slightly larger than the supply voltage -- the current ceases after 0.6· time-constant. The fraction of the inductively store energy that enters the bias battery may be determined as follows

$$\text{Energy to battery} = -\frac{\int_0^{t_1} E\, i\,dt}{1/2\, L\, I_o^2} = \frac{\int_0^{t_1} V\left[\frac{E}{R} - \left(\frac{E}{R} + I_o\right) e^{-(R/L)t}\right] d\cdot}{1/2\, L\, I_o^2}$$

$$\frac{-2E^2t_1}{RL\, I_o^2} - 2\left[\frac{E^2}{R^2 I_o^2} + \frac{E}{I_oR}\right] e^{-\frac{R}{L}t_1} + 2\frac{E^2}{R^2 I_o^2} + \frac{2E}{R\, I_o} = \qquad (3\text{-}5\cdot$$

$$\frac{2E^2}{I_o^2R^2}\left[\frac{I_oR}{E} - \ln\left(1 + \frac{I_oR}{E}\right)\right] = \frac{2}{\left(\frac{I_oR}{E}\right)^2}\left[\frac{I_oR}{E} - \ln\left(1 + \frac{I_\cdot}{E}\right.\right.$$

This ratio is shown on figure 3-114. As $\frac{I_oR}{E}$ is decreased to zero (accomplished by increasing the bias voltage with respect to the IR product), the fraction of the energy absorbed by the bias battery increases toward unity, and the portion dissipated

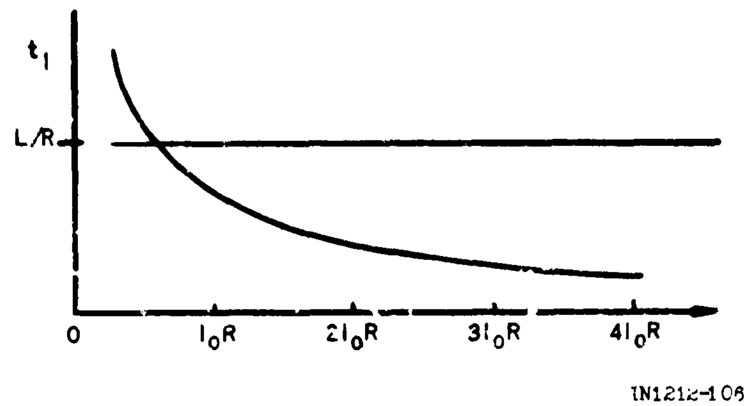


Figure 3-113. Current Decay Time vs Bias Voltage

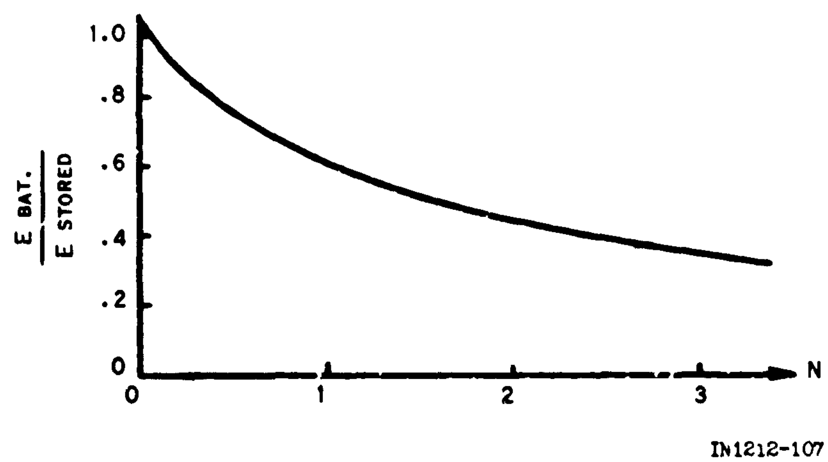


Figure 3-114. Energy Absorbed in Bias Battery vs N

in the load resistance and the rectifier decreases towards zero. Because the function of such an arc reducer is to prevent the occurence of large transient voltages across the switch contacts upon opening, and because the effectiveness of this limiting is dependent on the forward resistance of the diode, it is necessary to consider this aspect of the diode as well as its energy-dissipating capabilities. Considering the circuit shown on figure 3-110C, the voltage that appears across the contacts immediately following the switch opening is the algebraic sum of the source voltage, the bias voltage, and the forward voltage drop of the diode. This sum must not exceed approximately 300 volts if gaseous discharges between the contacts are to be avoided. To minimize this abrupt rise and the resulting interference, a rectifier having a low forward resistance should be used, and the bias battery eliminated.

(4) Composite Interference Reduction Components. Interference reduction can be obtained by using simple components in combination. Two composite arrangements are shown on figure 3-115. As each of the simple suppressors has some deficiency, it may be desirable to consider whether, when used together, each can supplement the other to improve over-all performance. It is usually easier to use an interference filter instead of a composite reduction circuit. To prevent the voltage across the switch from having an excessive peak value during opening of the contacts, the capacitor may have to be large. The required value of capacitance can be calculated from the circuit parameters. If, for instance, the load current is 3 amperes, the load inductance is 10 henries, and 300 is the maximum allowable voltage, then the

calculated value of capacitance is:

$$C = L(I/V)^2 = 10(3/300)^2 = 10^{-3} \text{ farads} = 1000\ \mu\text{fd} \quad (3\text{-}55)$$

The value of the capacitance (C) in composite circuit A on figure 3-115 is also a factor in determining the low-voltage breakdown or bridging. For larger capacitance values, the voltage build-up across the contacts upon switch opening is slower, and consequently, the electric field intensity within the gap is less. This consideration is a second factor in the selection of the value of this series capacitor. Under some circuit conditions, and with a fast opening switch, the value so determined will be less than the capacitance required to limit the peak switch voltage to a value lower than the breakdown potential of the surrounding gases. The procedures for determining resistance and diode characteristics are the same as described. The component arrangement of circuit A minimizes the formation of bridges, but is incapable of reducing sawteeth; the component arrangement of circuit B all but eliminates the sawteeth. Composite circuit B is a modified configuration. Here, circuit A is designed to minimize bridging, and circuit C is used on the sawteeth. Table 3-5 compares the ability of several circuits to retard the build-up of gap voltage, limit the peak of gap voltage, and minimize sharp wave-front transients. Table 3-6 compares the general interference reduction characteristics of components. The factors affecting the determination of component values are summarized in table 3-7.

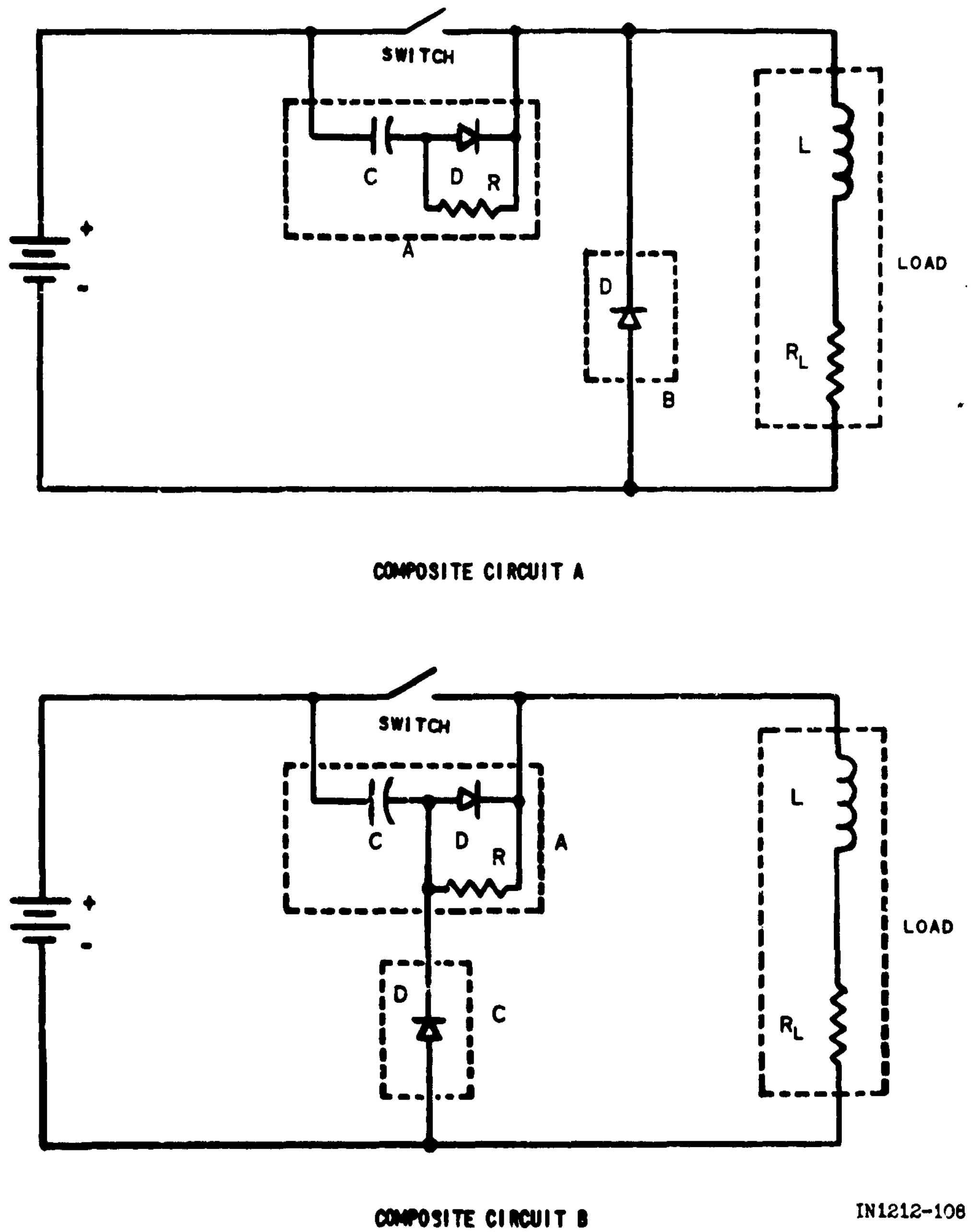


Figure 3-115. Composite Interference Reduction Components

TABLE 3-6. GENERAL INTERFERENCE REDUCTION CHARACTERISTICS OF COMPONENTS

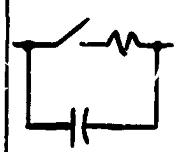
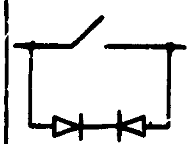
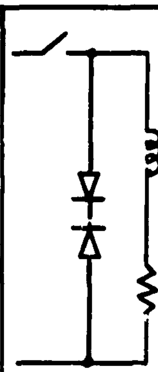
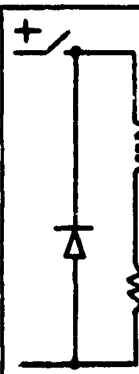
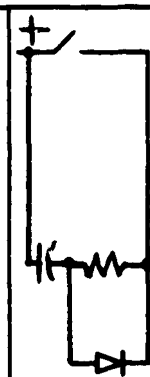
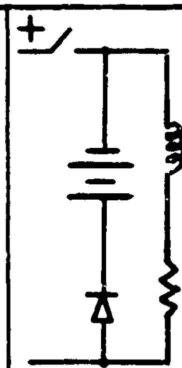
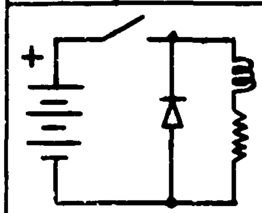
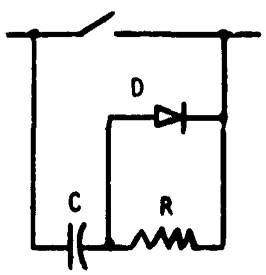
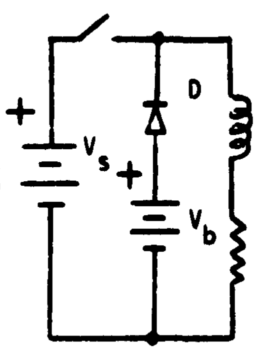
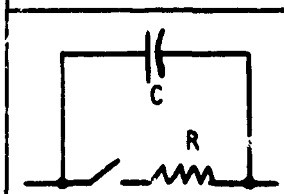
	Circuit					
Suppressor Feature						
Maximum Interference Reduction	Best	Good	Good to Poor	Poor	Good	Poor
Minimum Decay Time	Poor	Good	Good	Very Good	Good	Best
Ease of Adding to Circuit	Fair	Best	Best	Good	Fair	Poor (Need for Battery)
Size of Components For Small $1/2\ LI^2$	Small	Small	Small	Small	Small	Depends On Bias Battery
Size of Components for Large $1/2\ LI^2$	Large	Moderate	Moderate	Small	Large	

TABLE 3-7. DETERMINATION OF COMPONENT VALUES

Suppressor Circuit		Supply Voltage (V_s)	Load Current (I)	Load Inductance (L)
	Across Switch	Voltage for reverse knee must exceed supply	Reversed diode must dissipate all of stored energy. Select on heat basis. If duty cycle is high, current rating may be less than normal rectifier rating Knee < 300 V	
	Across Load	Same	Same	Knee < $(300-V_s)$
		Same	Diode must dissipate only a fraction of stored energy. Diode with rectifier, average-current rating equal to load current is conservative. If duty cycle is low, normal rating of diode may be increased by a factor up to ten.	
	D	No effect	Should have low forward resistance $R_f \leq = 1/I$	In practice, no effect. Diode, selected on rf basis, could carry load current indefinitely
	C	No effect	$C \leq L(I/300)^2$ to eliminate sawteeth	
	R	No effect	Must completely discharge C during switch-closed interval. R = T*/5C	
	D	Voltage for reverse knee must exceed: $V_b + V_s$	R_f must be less than: $\frac{300 - V_b - V_s}{I}$	If V_b is large (to reduce decay time), diode selected on basis of rf could carry load current indefinitely. No effect of L
	V_b	If "slugging" is to be minimized, then V_b should be maximum possible and still keep gap voltage less than 300 V: $V_s + V_b + IR_f < 300$ V		
	C	No effect	$C \geq L(I/300)^2$ to eliminate sawteeth	
	R	No effect	R = value which produces maximum allowable voltage regulation	No effect

* T = duration of closed interval

3-22. Electron Tube Control Circuits

a. Vacuum or gas-filled electron tubes are often used to produce switching actions or to control switching circuits. In some applications, the generation of harmonics is desired, or essential, to the operation of the power or control system. Harmonic generation, however, also gives rise to interference.

b. Electron tubes are generators of nonsinusoidal waveforms and are widely used as pulse generators, modulators, or oscillators in chopper circuits. The greater the power output of these control devices, the more difficult it is to keep the harmonic content and interference under control. If gas-filled diode tubes are used, interference increases as tube conduction increases. Less interference and more flexible control may be achieved by using a vacuum triode and applying a control voltage to the grid instead of relying entirely on the plate voltage to cause breakdown (switching action). Gas tubes usually permit more precise conduction control.

c. In control systems using thyratrons, the interference source is usually located in the tube. Inserting filters in the thyratron plate circuits, as close to the plate terminal as possible, will reduce conducted interference. A choke coil of approximately 15 millihenries, with a current-carrying capacity comparable to that of the tube, preceded by a capacitor (0.1 microfarad) to bypass radio frequency to ground, can provide up to 40-db attenuation below 50 mc. To minimize radiation effects, the thyratron should be physically isolated from magnetic field effects from other components and should be electrostatically shielded to reduce spurious radiation from mercury vapor discharge. Interference, arising in mercury-arc electron tubes, characteristically consists of oscillations of several hundred kc superimposed on a random interference background extending up to about 5 mc.

3-23. Magnetic Amplifiers

a. A magnetic amplifier controls the reactance of a coil by controlling the effective permeability of its magnetic core material. In its simplest application, the magnetic amplifier is inserted in series with the load impedance of a circuit. As the impedance of the amplifier is varied by changing the degree of saturation of the core (with a small change in dc, or properly phased ac current in a separate winding on the same core), the power to the load is increased or decreased. An unsaturated core presents a relatively high impedance to ac; but a saturated core acts as an air core with practically no impedance. There are two distinct advantages to be gained by using magnetic amplifiers as switching-circuit control devices: Gain control leads are less affected by stray rf pick-up than those of equivalent electron-tube circuits, and signals, that are too weak to override shot noise in vacuum tubes, can be amplified.

b. Magnetic amplifiers should be considered as possible sources of interference. The harmonic content of the output of some magnetic amplifiers is one possible interference source. The harmonic content depends somewhat on the amount of relative control used (net ampere-turns of control signal applied to the control windings). The waveform itself can generate interference under certain conditions. For example, a load connected through brushes and a commutator can cause steep wave-fronts and high-frequency interference.

c. The amplistat, a self-saturating form of the static magnetic amplifier, has a pulsating dc output. If the amplistat load is inductive, the current peaks should be smoothed out, but voltage peaks could puncture the rectifiers used in the device. Such induced voltage peaks produce interference and should therefore be controlled by using such circuits as diode rectifiers. Inductive loads also cause core flux unbalance and voltage to be induced in the control windings. Induced voltage causes considerable double-frequency current to flow if the control circuit impedance is low. The current will then add to, or subtract from, the operating control current and influence the operation of the amplifier. With low-circuit impedance, the output voltage will be unstable in the low-voltage region; that is, for certain values of apparent control ampere-turns, the output voltage snaps up or down depending on whether the control current is increasing or decreasing. Such snapping can generate interference. To eliminate snapping, a large control circuit impedance should be provided, or interference reduction circuits should be incorporated at the amplifier load.

3-24. Transistor Switching Circuits

a. Transistor circuits can sometimes be used advantageously to reduce switching circuit interference. Gaseous discharge and bridging interference at switch contacts can be eliminated by reducing the switch current to that required for transistor operation. Transistors also offer advantages in reliability, size, and weight for many applications. The design engineer should therefore carefully consider their use rather than depend solely upon resistors, capacitors, inductors, or rectifiers. Typical transistor switching circuits are shown on figure 3-116. Base-to-emitter potential is an important consideration when using transistors in these circuits. The figure shows that a controlling set of contacts may be inserted in the base lead, or across the base and emitter terminals, of a common-emitter transistor circuit. Because the floating potential of the base is about 0.1 volt (V_{be} with the base open), almost the entire base-to-emitter supply voltage will appear across the contacts if they are inserted in the base lead. If an auxiliary power supply is not available and relatively high voltages are to be switched, it may be advisable to place the switching contacts across the base and emitter terminals. In this case, the voltage across the contacts will be the emitter-to-base forward voltage (0.5 to 3 volts).

b. Transistor circuits reduce interference in inductive circuits by limiting the peak induced voltage when the switching circuit current is interrupted. The peak induced voltage usually far exceeds the back-voltage ratings of the base-to-collector diode of most transistors and causes an effect known as punch-through, or reverse breakdown. If the power limits of the transistor are not exceeded, no damage will result from punch-through. Furthermore, the effect is advantageous because di/dt is reduced, and the induced voltage, -L(di/dt), is also reduced. To take full advantage of punch-through, a transistor with a breakdown

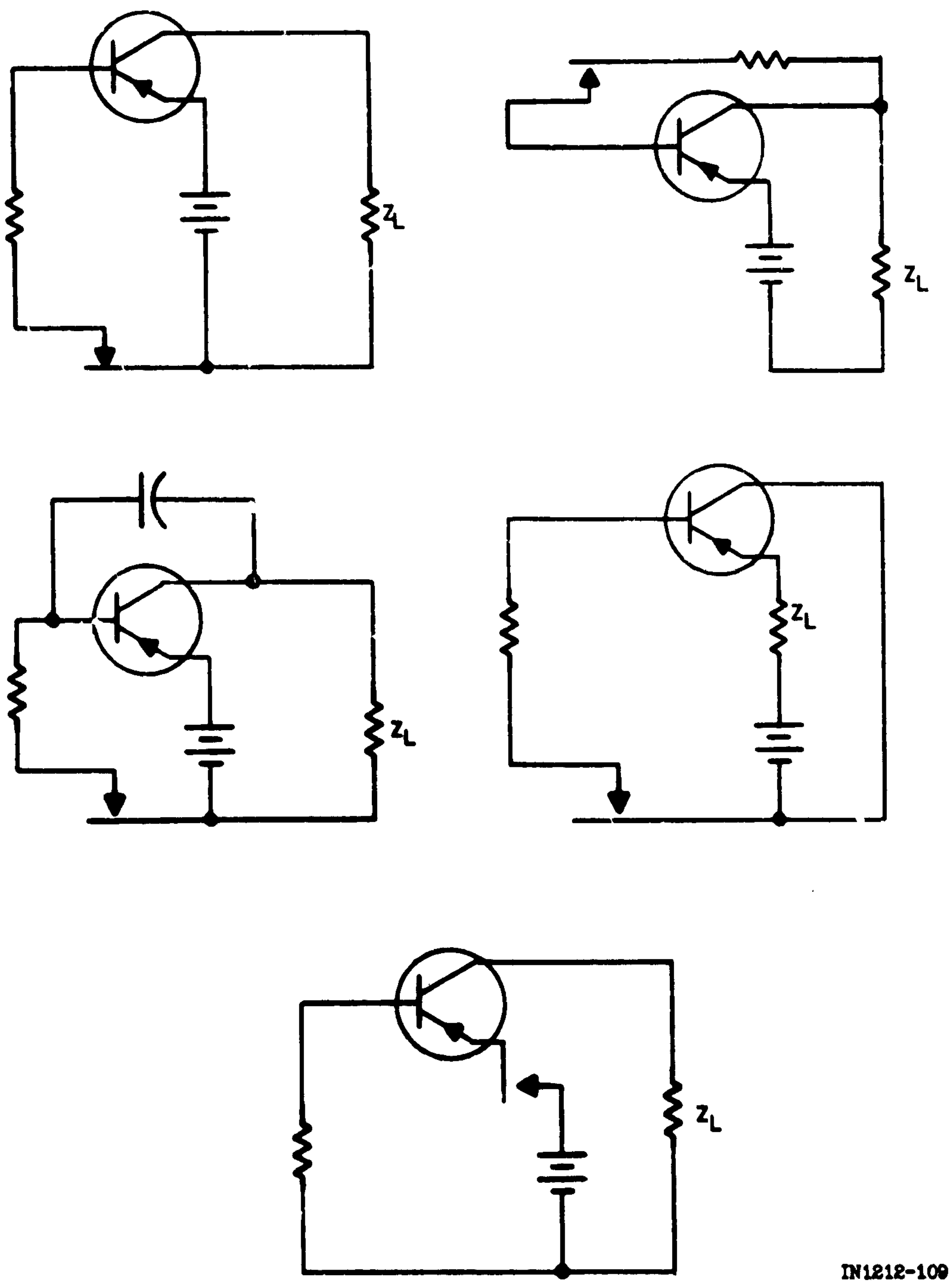


Figure 3-116. Transistorized Control Circuits Designed For Interference Reduction

voltage nearly equal to the source voltage should be used. High-frequency interference, generated at the load side of the transistor, is then effectively reduced. Interference can be further reduced when a low-impedance feedback network is used with a common emitter circuit, and the load impedance is not less than the base resistance of the transistor.

3-25. Filtering

The classic technique for the reduction of interference generated by switches is the use of a filter network, consisting of one or more inductors and one or more capacitors, all mounted within a grounded metal shield (fig. 3-117).

a. Series Inductance and Shunt Capacitance in Combination with Resistance Capacitance. Low-pass filters, consisting of series inductance and shunt capacitance, are effective in reducing interference caused by switch transients (fig. 3-118). The low-pass network is effective but has the disadvantage that the surge is not completely eliminated because of the distributed capacity of the inductances and the inherent inductance of the capacitor leads. Using a shielded enclosure and feed-through capacitors avoids these difficulties.

b. Shielded Enclosure and Feed-through Capacitors. When feed-through capacitors are added to the circuit shown on figure 3-118, the inherent inductance from line to ground is reduced to an almost negligible value (fig. 3-119). The resonant frequency of the capacitor also is raised, and its usefulness for effective bypassing is increased.

c. Conventional Double-Pi Filter. An arrangement such as the one shown on figure 3-120A is ofte effective in reducing interference in an inductive circuit. The extent of reduction depends upon the nature of the interference and the capabilities of the filter. The filter's insertion-loss curve should be consulted when a particular filter is required to attenuate undesirable interference.

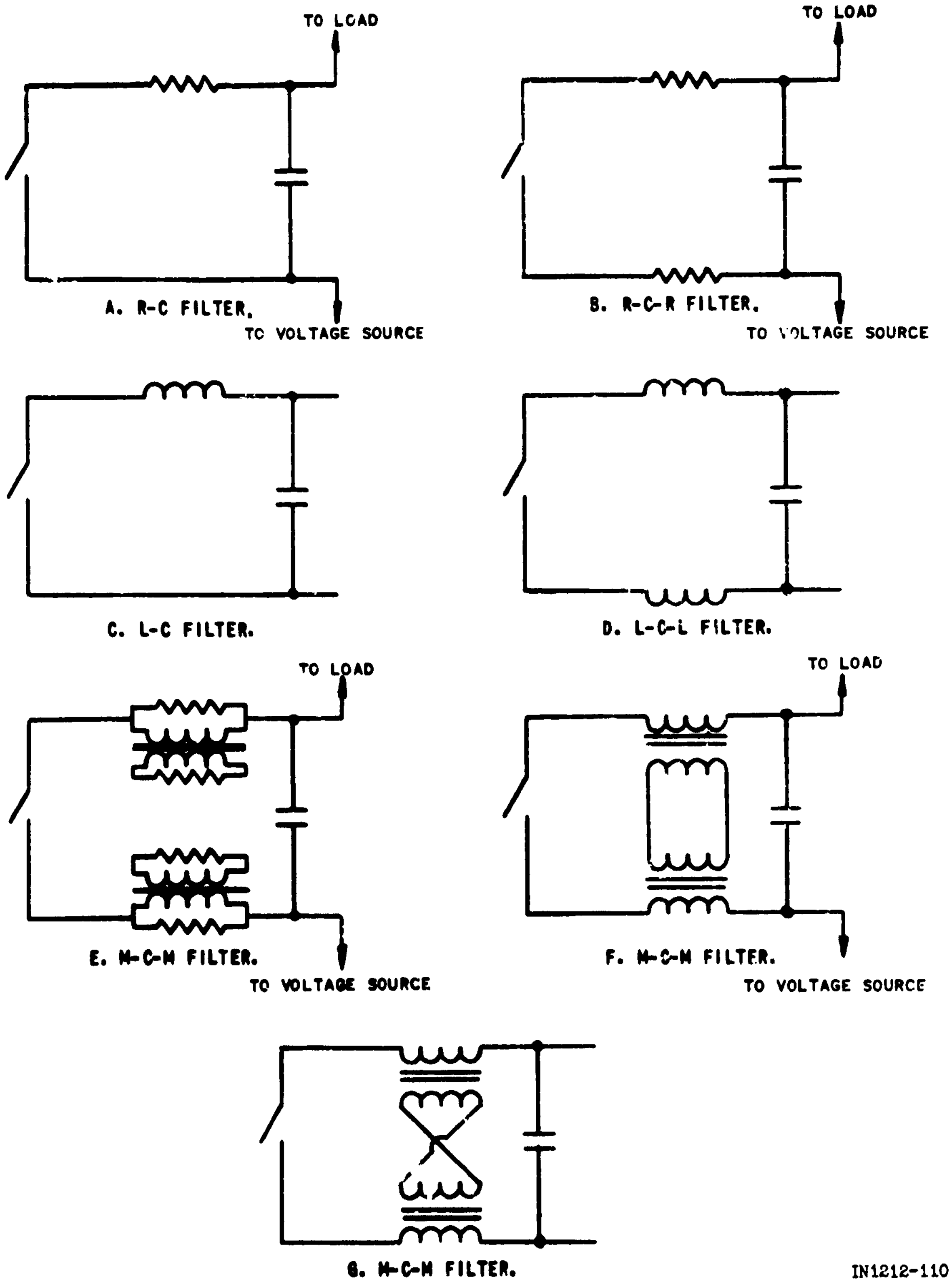


Figure 3-117. Switch Filter Networks

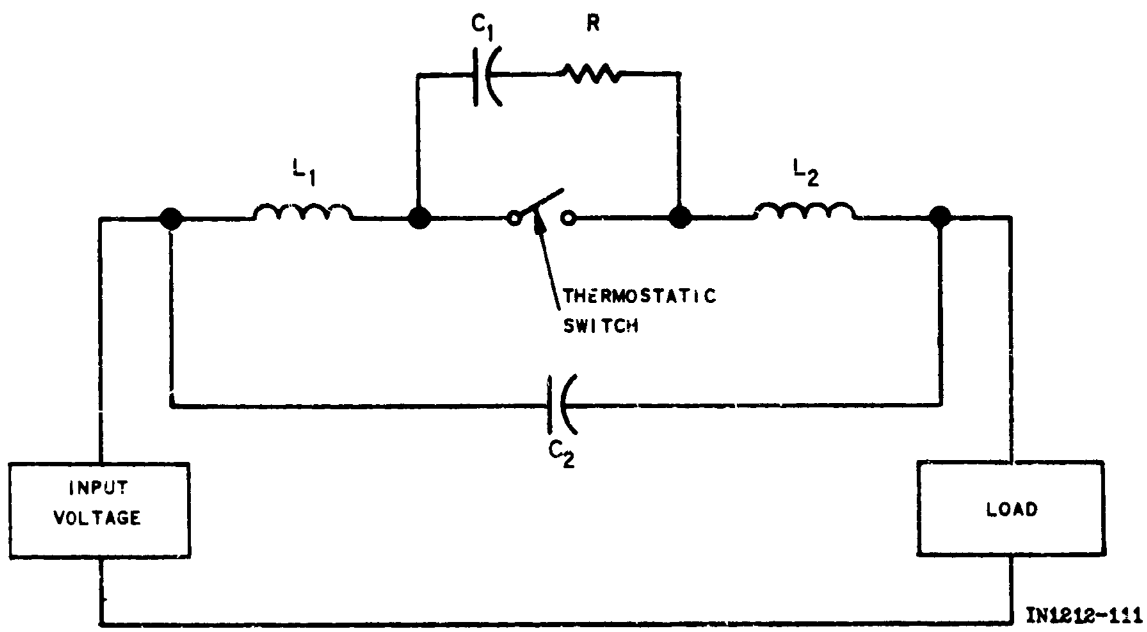


Figure 3-118. Composite Interferance Reduction Filter

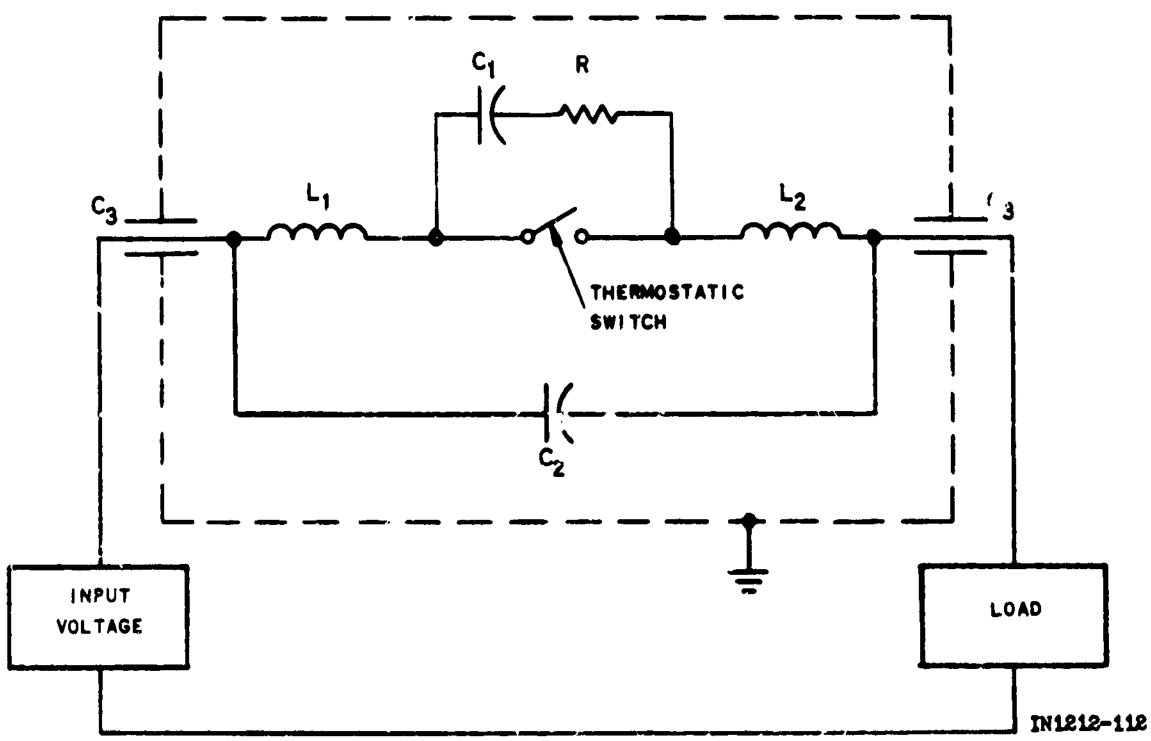


Figure 3-119. Interference Filter in Shielded Enclosures

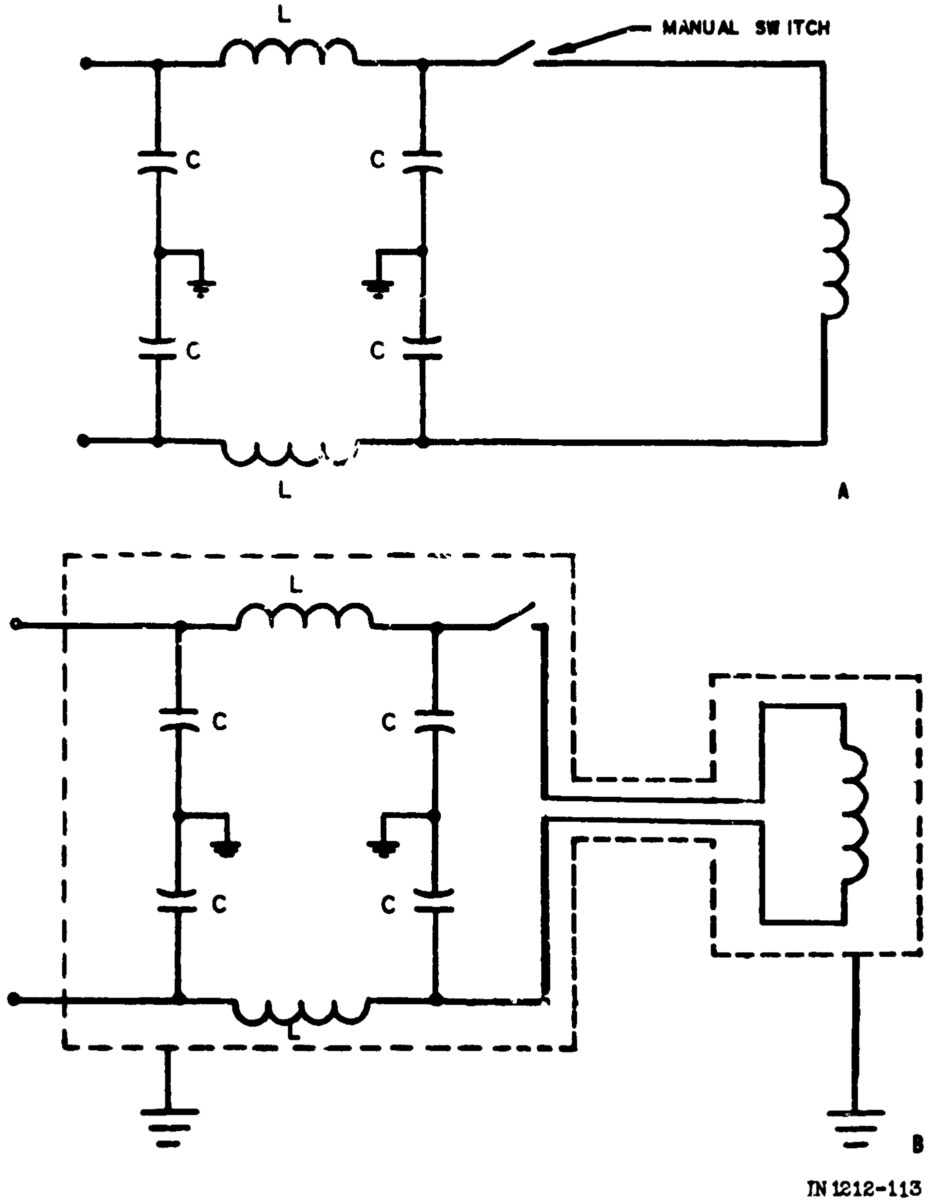
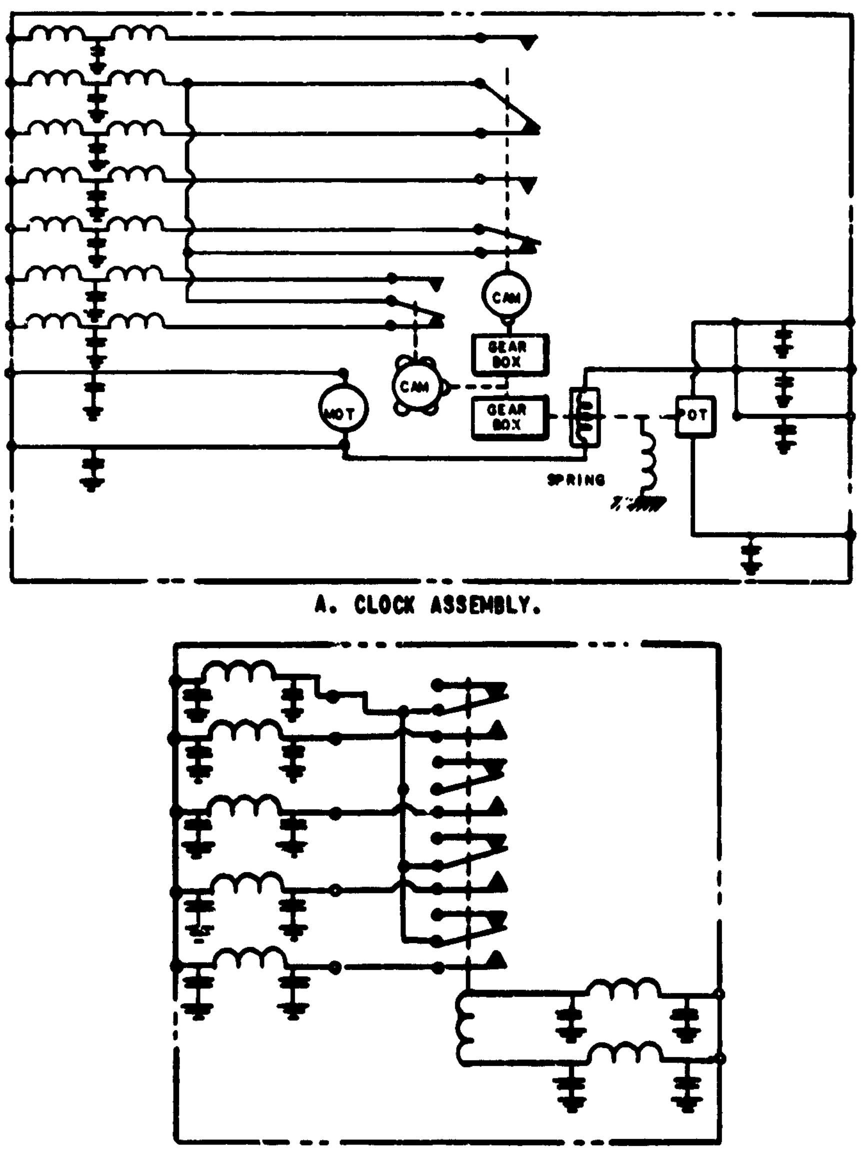


Figure 3-120. Double-PI Filter

d. Filter, Shielded Leads, and Shielded Switch. An arrangement (fig. 3-120B) that employs shielded conductors, a filter, and a shielded enclosure is effective in reducing interference in even the most difficult cases. Where a great many switches are used, it is often advantageous to first filter all leads entering the package and then shield the package (fig. 3-121). See section 2 of this chapter for a detailed discussion of filter networks.



A. CLOCK ASSEMBLY.

B. TRANSFER RELAY.

IN1212-114

Figure 3-121. Application of Filtering and Shielding

Section V. RECEIVERS

3-26. General

Basically, interference problems in receivers may be reduced to two major areas: (1) undesired responses and (2) undesired emissions. Undesired responses encompass two general areas or characteristics: susceptibility and spurious responses. Undesired emissions include unwanted signals or interference generated within, and propagated from, the receiver.

a. Susceptibility is normally defined as the characteristic of a receiver to respond to signals entering through paths other than through its antenna circuit. Spurious responses are defined as the response of the receiver to signals across its antenna terminals at frequencies outside its pass-band and are usually expressed in terms of db above the receiver's normal sensitivity. While much can be done and is being done in military equipments and systems to reduce interference at its source, operational requirements frequently make it necessary for receivers to operate in the presence of high level signals from transmitters in the immediate vicinity. Therefore, the susceptibility and spurious response characteristics of a receiver are of vital importance if its operational purpose and capability are to be maintained.

b. The spurious response and susceptibility characteristics of a receiver are affected by operating frequency, linearity, type of circuits, (ie: superhet, crystal video, etc.), bandwidth, sensitivity and other factors. Physical design and layout, particularl, in the "front end", are a most important consideration. Coupling between internal and external circuits, as affected by placement of components, isolation between wiring and circuitry, circuit shielding, decoupling and filtering, are major factors relative to susceptibility. There are four basic paths through which an undesired signal can enter a receiver. They are illustrated in simplified form in figure 3-122 as follows:

1) The antenna system
2) Power and control circuits
3) Interconnecting and output circuits
4) Case penetration

Conversely, interference generated within a receiver may be propagated through these same paths; and many of the measures and techniques which minimize spurious responses and susceptibility of a receiver will also serve to attenuate or contain interference generated within the receiver. Interference generated within a receiver must also be minimized, or contained, to prevent interfering with other nearby receivers and, in tactical operations, to avoid detection by the enemy. Under favorable conditions, receiver local oscillator radiation may be detected several miles away.

c. Many interference and susceptibility problems in receivers could be readily avoided in the design, if only some thought is given to the fact that the signals are not confined to the signal paths indicated in the schematic diagram. It is necessary to design into the receiver the physical confinement of both the internal and external signals. The representation of figure 3-122 illustrates, in the very simple case, various paths by which undesired signals may enter a receiver. These same deficiencies permit propagation of local oscillator, and if signal energy out of the receiver. Figure 3-123 similarly illustrates certain basic interference reduction measures applied to a simple receiver.

d. In analyzing a receiver design, each component, each circuit, and each lead must be viewed with suspicion as to its capability to propagate or accept unwanted signals. Having this concept or philosophy in mind, the designer will often detect and resolve many potential interference problems. Interference problems in receivers, which must be taken into account in the basic design as well as in actual layout and fabrication, may be summarized as follows:

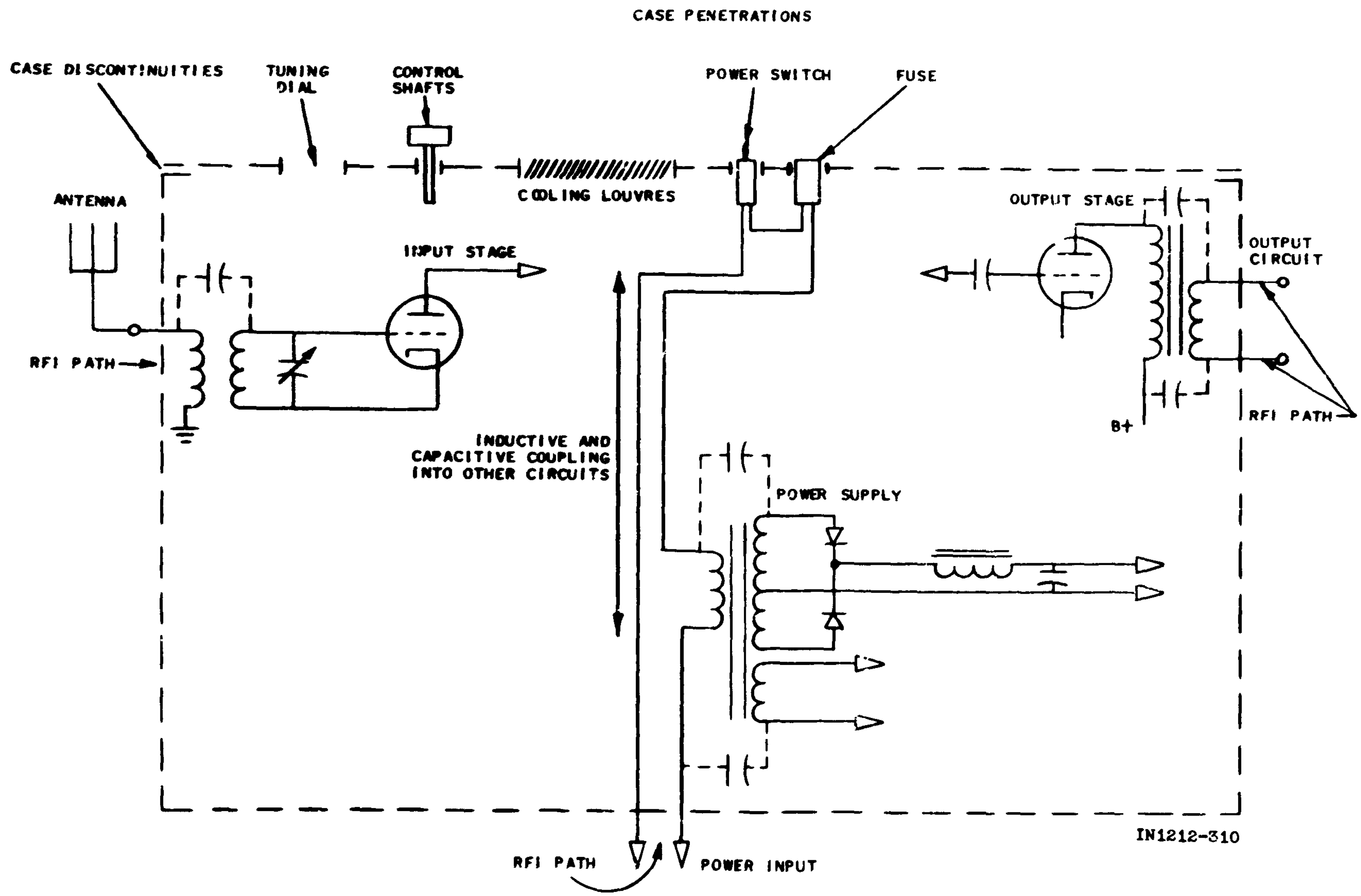


Figure 3-122. RFI Paths Into Receiver

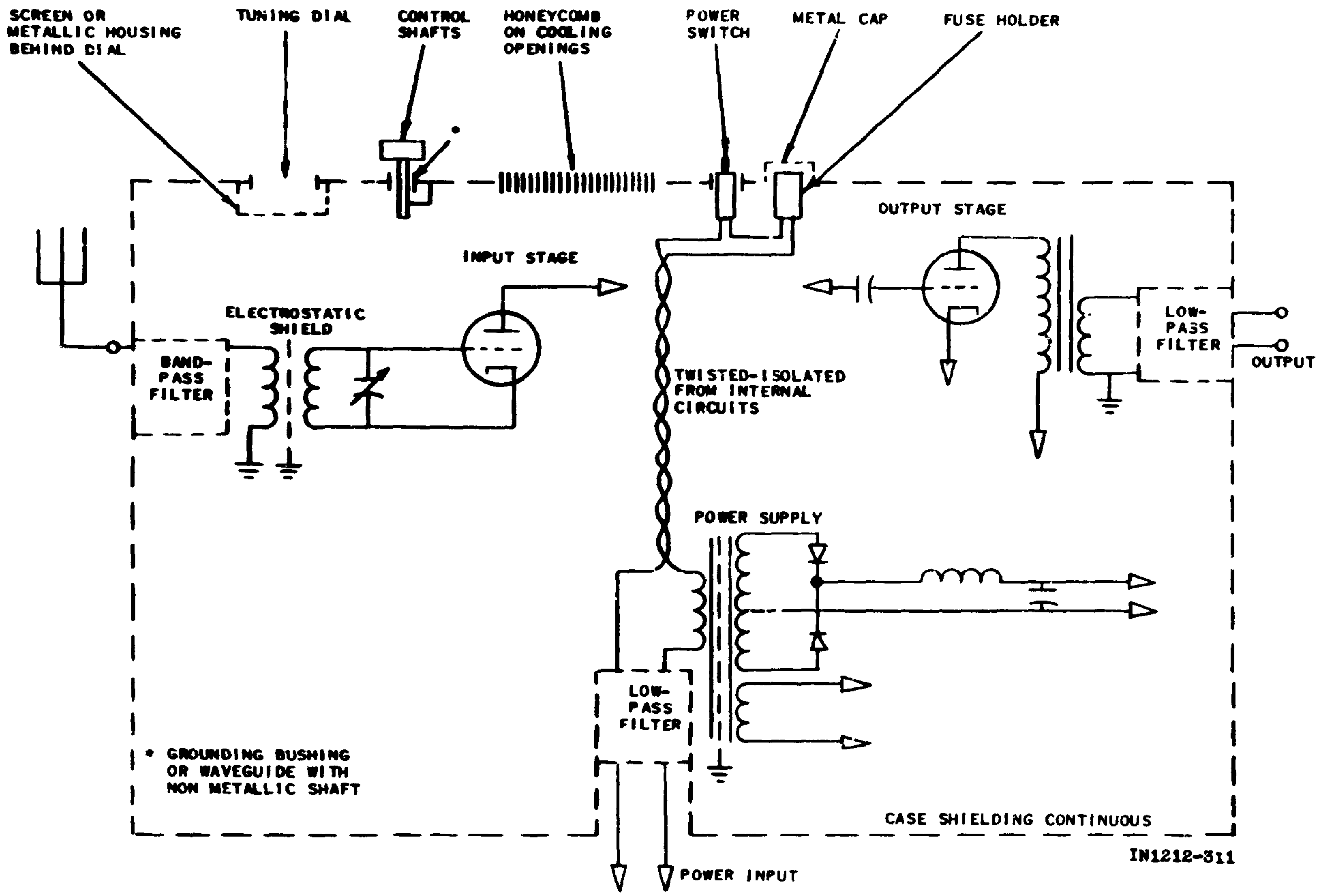


Figure 3-123. Basic RFI Suppression of Receiver

(1) Interference generation

(a) Broadband interference generated by solid-state power supplies, relays, and other switching devices. (Suppression techniques for these devices are covered in other sections of this manual)

(b) Video signal radiation and conduction via power and other lines

(c) Local oscillator, fundamental and harmonics, radiating from antenna, housing or wiring

(d) Intermediate frequency signals and harmonics

(2) Susceptibility

(a) Conducted signals at rf, if, or image frequencies through power circuits, signal output circuits, control or other circuits entering or exiting receiver housing

(b) Radiated susceptibility, due to shielding deficiencies allowing signals to enter receiver housing, shielded interconnecting cables, connectors, etc., and reaching low-level circuits

(c) Audio susceptibility (due to audio ripple on dc lines causing audio response), instability or modulation of rf, if or video signals, or mixing with or producing command signals. This is a particularly serious problem in transistorized receivers or portions thereof, usually due to inadequate low frequency decoupling

(3) Spurious responses

(a) Image response

(b) IF response

(c) Intermodulation

(d) Cross modulation

The broad category of spurious responses covers a wide variety of undesired responses to signals outside the receiver pass-band as a result of normal and abnormal non-linearities, inadequate front-end selectivity, and/or spurious resonances. Receiver desensitization can also be considered within the category of spurious responses, since it is generally a result of a strong signal outside the receiver pass-band causing overloading of the rf or mixer stage, or reduction of gain by agc action.

3-27. General Interference Reduction Methods

Basically, interference reduction in a receiver is accomplished by attenuating the propagation of undesired rf energy through any of the typical paths indicated in figure 3-122. The basic techniques employed include:

1) Filtering
2) Shielding
3) Isolation of internal wiring and circuit components
4) Internal decoupling and filtering
5) Circuit design and layout

The techniques which are effective in reducing susceptibility of a receiver are similarly effective in reducing interference generated within the receiver. Frequently, the same deficiencies which permit the entry of undesired rf energy into the sensitive circuits allow radiation or conduction of local oscillator or if signals out of the receiver. Inasmuch as the rf environment in which a receiver is to be employed cannot generally be determined by the designer, permissible levels of interference emanation and susceptibility are specified in the applicable MIL interference specification or in the equipment specification.

<u>a.</u> <u>Filtering.</u> In other than rf signal circuits, such as power input, output, or control circuits, simple low-pass filters are most commonly employed. In such circuits, where it is necessary or desirable to atten-

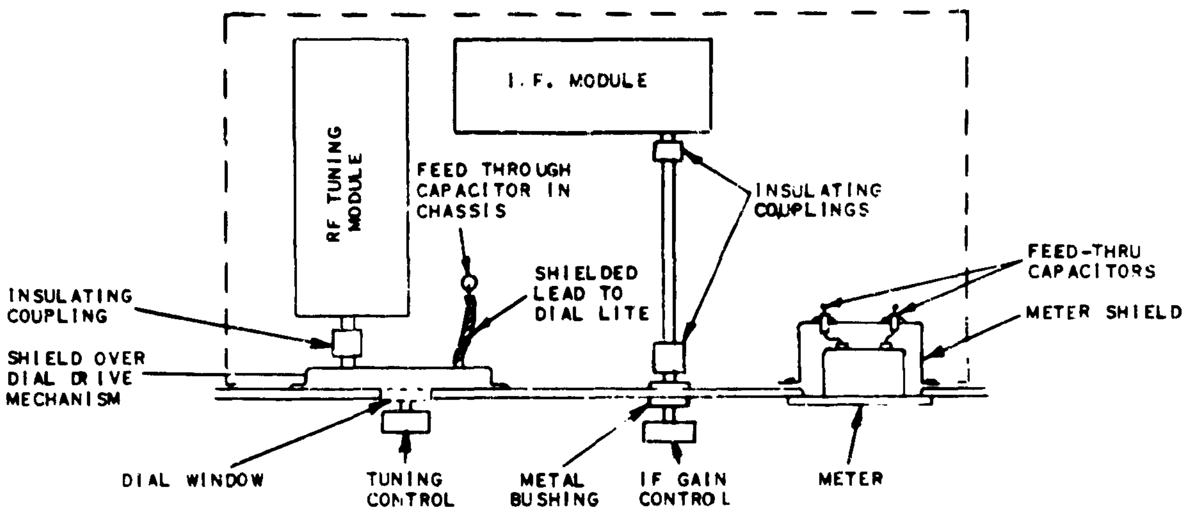
uate rf energy over a wide frequency range, low-pass filters should be used. The cutoff frequency is limited only by the power frequency, audio, digital, or other information which must be passed. In rf circuits such as the rf input, more complex types of low-pass, band-pass or high-pass filters may be employed depending upon the tuning range and other factors. Tunable band-pass filters are sometimes employed, but these become rather bulky and require manual tuning. Such filters are usually employed only to overcome internal receiver deficiencies, or where a receiver must operate in close proximity to high-power transmitters where rf power levels exceed practical rejection capabilities.

(1) Receiver rf input. In the hf, vhf, and uhf ranges, receivers are commonly of the superheterodyne type with tuned rf circuits. The tuned circuits themselves are basically band-pass filters. In some receivers, the number and Q-values of these tuned circuits are inadequate to provide sufficient off-channel rejection of strong signals. For such cases, one or more additional tuned circuits may be used at the input, preferably isolated by additional rf amplifiers. The effectiveness of the receiver tuned rf stages as a band-pass filter can be seriously degraded by inadequate isolation and/or decoupling between the successive tuned stages. This is a common problem frequently encountered in otherwise well-designed receivers. Coupling around tuned circuits via filament, B+ or agc circuits, or the lack of proper shielding between stages, are among the major causes of inadequate rejection of high-level signals outside the receiver pass-band. For reducing interference outside the normal tuning range in variable or in fixed-tuned receivers, band-pass and low-pass filters in the antenna circuit provide a relatively inexpensive approach. In fixed-tuned receivers, the pass-band of the filter(s) can be made sufficiently narrow to also be effective in reducing local oscillator signal feed-

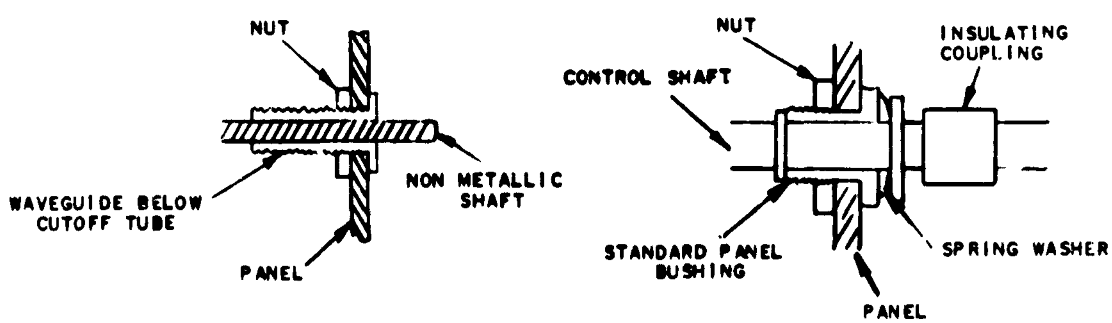
back into the antenna. IF rejection can be improved by incorporating high-pass and low-pass filters in the antenna circuit, but the use of low-pass filters to improve if rejection is considered more of a fix than a desirable design feature. In the basic design, rf selectivity through the use of well-designed tuned circuits and reduction of stray coupling around the tuned rf stages is the proper approach; untuned rf stages are undesirable. In some cases, even the best designed tuned rf preselector can fail to offer sufficient rejection of signals of much higher frequencies; it may then be necessary to add a low-pass filter which has its cut-off frequency where the tuned circuit starts to fail.

b. Shielding. Shielding within a receiver is an important consideration which should be undertaken in the initial design. The addition of shielding into an existing receiver is almost totally impractical and at best, a compromise. Good interference reduction design requires shielding in several forms which may be divided into two primary areas, external and internal. The purpose of shielding, stated simply, is to contain or exclude unwanted rf energy. In a receiver, both of these requirements exist: (1) to contain signals generated within the receiver and (2) to exclude high level rf energy which, by entering the receiver through the "back door", is not rejected by the rf preselection. Obviously, if such signals are at an if frequency, a response will be produced. However, frequency coincidence is not necessary since responses may also be produced, for example, by mixing with the local oscillator or its harmonics. Shielding and decoupling, or filtering, are interdependent and either may enhance or degrade the effectiveness of the other. However, as mentioned earlier, filters for decoupling may be much more readily added to a completed well-shielded receiver or circuit than shielding, which usually requires major rework. Therefore, shielding of the receiver and critical circuitry should be considered from the outset of design and layout. It must also be remembered that the degree of internal shielding required to merely permit a receiver to operate without oscillation or unstability is not normally adequate to meet rfi requirements.

(1) Housings. Figure 3-124 illustrates in simple form a few of the basic shielding considerations applicable to the overall receiver housing. Obviously, any receiver case or housing must provide access for maintenance or removal of the receiver chassis. However, removable panels should be minimized insofar as practicable. Continuously welded or closely spaced spot-welded construction of the basic housing should be used wherever possible. Access panels should be attached by machine screws, with continuous metal-to-metal contact between the mating surfaces. Where continuous metal-to-metal contact cannot be maintained around the entire periphery of the opening, due to surface irregularities or deformation of the metal, rf gasketing should be used. Types and typical applications of rf gasketing materials are described in Section IV, Shielding, of Chapter 2. In considering the practical aspects of overall receiver shielding, the discontinuities and penetrations of the housing are usually the limiting factors, rather than the type or thickness of metal employed. Normally in a receiver housing, either aluminum or sheet steel of sufficient thickness to provide the necessary mechanical strength will provide adequate shielding, but the mating surfaces and openings are the major problem areas. Where pressure-sealing is required, such as in airborne, missile or certain drone applications, rubber or synthetic "O" ring seals may be used between sections of the casting. Care should be taken either to insure that the "O" ring is completely compressed in a groove to allow metal-to-metal contact between sections of the casting, or a second groove should be provided to accomodate a woven metallic mesh gasket for rf bonding. Cast housings of magnesium should be avoided



A. SHIELDING CONSIDERATIONS FOR CONTROL PANEL PENETRATIONS



B. DETAILS OF TYPICAL SHAFT PENETRATIONS IN1212-312

Figure 3-124. Receiver RFI Suppression

because of corrosion of the mating surfaces. Aluminum castings having machined mating surfaces should be protected by plating or other finishes such as Iridite which does not appreciably impair the conductivity. Electroless nickel plating provides an extremely good finish for aluminum. It prevents oxidation and corrosion and maintains excellent rf properties. Except for special purpose fixed-tuned receivers, openings for controls and indicators are necessary. These and cooling or ventilation openings represent the major areas of rf leakage in the average receiver housing. No control shaft should be continuous from the external control into an rf module or circuit. Tuning control shafts, for example, should be "broken" by an insulating coupling. At the point of penetration of the control panel, the control shaft should be grounded to chassis by a suitable bushing. A non-conducting shaft penetrating the housing through a metal tube having at least a 3:1 ratio of length to diameter, is another means of treating control penetrations to maintain shielding integrity. Openings for cooling or forced air ventilation must be shielded, either by the use of fine mesh screen or honeycomb-type filters. The latter type offers several advantages over screening in that:

1) It offers little impedance to air flow
2) It affords generally greater shielding effectiveness
3) Clogging, such as often occurs with fine mesh screening, is not a problem

It does, however, require somewhat more space since its attenuation characteristics are a function of the depth relative to the cross-section of the individual openings. In either case, complete rf bonding around the entire periphery of the opening must be provided, whether screen or honeycomb material is used. This bonding is best accomplished by continuous soldering to the housing or enclosure around the opening. In the case of aluminum enclosures, the screen or honeycomb material may be soldered into a frame which is, in turn, mounted in the enclosure.

(2) Internal shielding. Internal shielding within a receiver is equally as important as the external shielding or receiver housing. Shielding of individual circuits or sections of a receiver, besides affecting its stability, affects spurious responses, susceptibility and spurious emissions. Internal shielding, as in the case of the overall housing, is also greatly dependent upon circuit decoupling and filtering.

(a) Ideally, each stage of a receiver from the antenna through the second detector should be housed in an individual shielded compartment, with only those conductors required to carry necessary signals running between the compartments. Power, agc circuits, etc., should be routed separately to each compartment and filtered at the point of entry into the compartment. A typical example of a shielded rf module (and methods of applying decoupling to power and/or control circuits) is shown in figure 3-125. The rf gasket is in the form of woven metal mesh inserted in a channel around the periphery of the module shield. The cover may be fastened by machine screws through the top (which pulls the cover down compressing it into the rf gasket). In the case of non-removable modules, rf shields may be applied in the same manner; but with the feed-thru rf filters mounted on the chassis.

(b) The importance of adequate grounding and bonding between sections of a shielded compartment, or between a shield and the chassis, cannot be over-emphasized. Continuous bonding must be maintained around the entire mating surfaces in order for a shield to be fully effective. This becomes more critical with increasing frequency and with high-level signals, such as in a local oscillator. The use of metallic mesh gasketing,

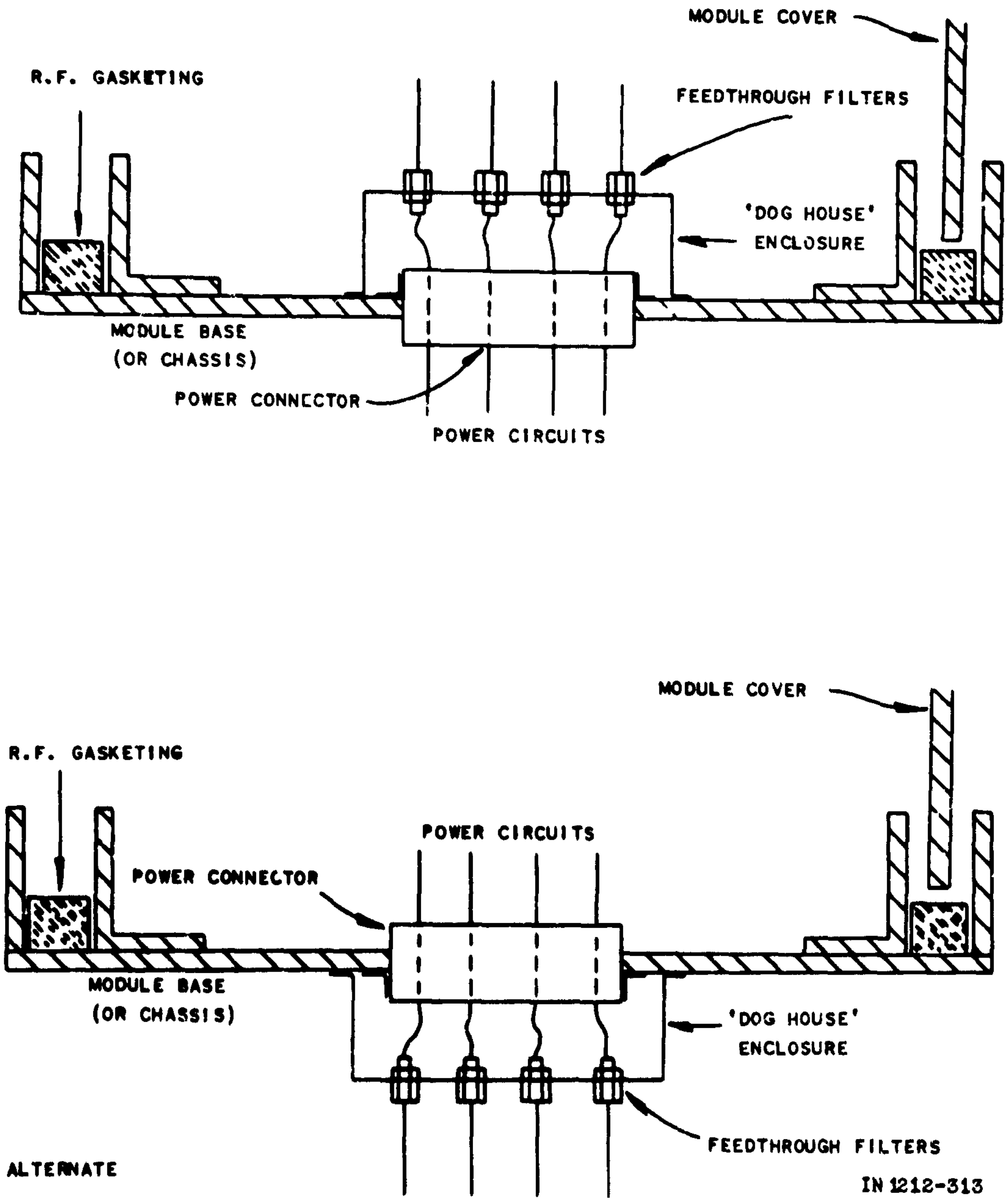


Figure 3-125. Module Base with Feedthrough Components Mounted In A "Dog-House" Enclosure

as shown in figure 3-125, is one means of insuring complete rf grounding of a shield housing or a cover without having to maintain extremely close tolerances on mating surfaces. This method has been proven effective so long as adequate pressure (approximately 20 psi) is maintained on the gasket.

(c) In variable tuned receivers where the rf, mixer, and oscillator tuning must be tracked, shielding between these stages becomes mechanically more difficult to accomplish. However, this shielding is critical with regard to oscillator radiation and spurious responses. As mentioned earlier, it is preferable to contain each of these in its own individual shielded compartment. The greater the departure from this ideal situation, the greater the compromise in interference and spurious response characteristics. Wherever practical, the mechanical linkage employed for tracking the various tuning elements should be electrically isolated between the various tuned stages. If the rf, mixer and oscillator tuning employs ganged tuning capacitors, or variable inductors utilizing a single metallic shaft, extreme care should be taken to provide low impedance grounds on the shaft between each stage. When such a tuning arrangement is used and incorporated in a single enclosure, shield barriers should be utilized between each stage. This shielding barrier should provide continuous electrical contact with the enclosure. Penetrations of the shields between stages should be held to the minimum required for interstage signal coupling and tuning linkage. Circuits should be so laid out as to minimize lengths of signal leads between stages. Local oscillator circuitry should be physically, as well as electrically, isolated from the rf stage(s). Coaxial cable, entering or exiting a shielded rf circuit or stage, should have its shield grounded at the point of penetration either by means of an rf connector or by connecting the shield direct to chassis. In an area of high rf fields, such as in an oscillator module, rf energy can be coupled out on the cable shield unless it is properly grounded at the point of entry. In the design of rf transformers, electrostatic shields should be employed between primary and secondary wherever practical. This reduces the capacitive coupling between windings, decreasing spurious coupling and resonances; thus, reducing spurious responses and coupling of the oscillator signal back to the antenna circuit.

(d) Test points, frequently required in military receivers, are often a source of interference and susceptibility problems. Where it is necessary to provide test points to monitor voltages or circuit functions in the rf or if portions of a receiver, care must be taken to provide proper decoupling and shielding of the leads to the test points. Where a panel is provided for a number of test points, it should be recessed and provided with a metallic cover to maintain overall shielding integrity. Where a test point for the oscillator signal is required, the oscillator signal should be rectified and filtered in the oscillator module, and the resulting dc monitored rather than the oscillator signal itself.

(e) Figure 3-126 shows a typical example of a poorly designed vhf "front end" wherein the rf stage, mixer and oscillator are contained in a single shielded unit. The unit exhibited not only inductive and capacitive coupling between the oscillator circuitry and the rf stage but, due to poor decoupling techniques, a direct path from the oscillator through C-1, R-1, R2 and C2, and back to the antenna input transformer, T-1. This particular receiver produced some 2200 microvolts into a 50-ohm load across the antenna terminals at the oscillator frequency, with high levels appearing through the sixth harmonic. Local oscil-

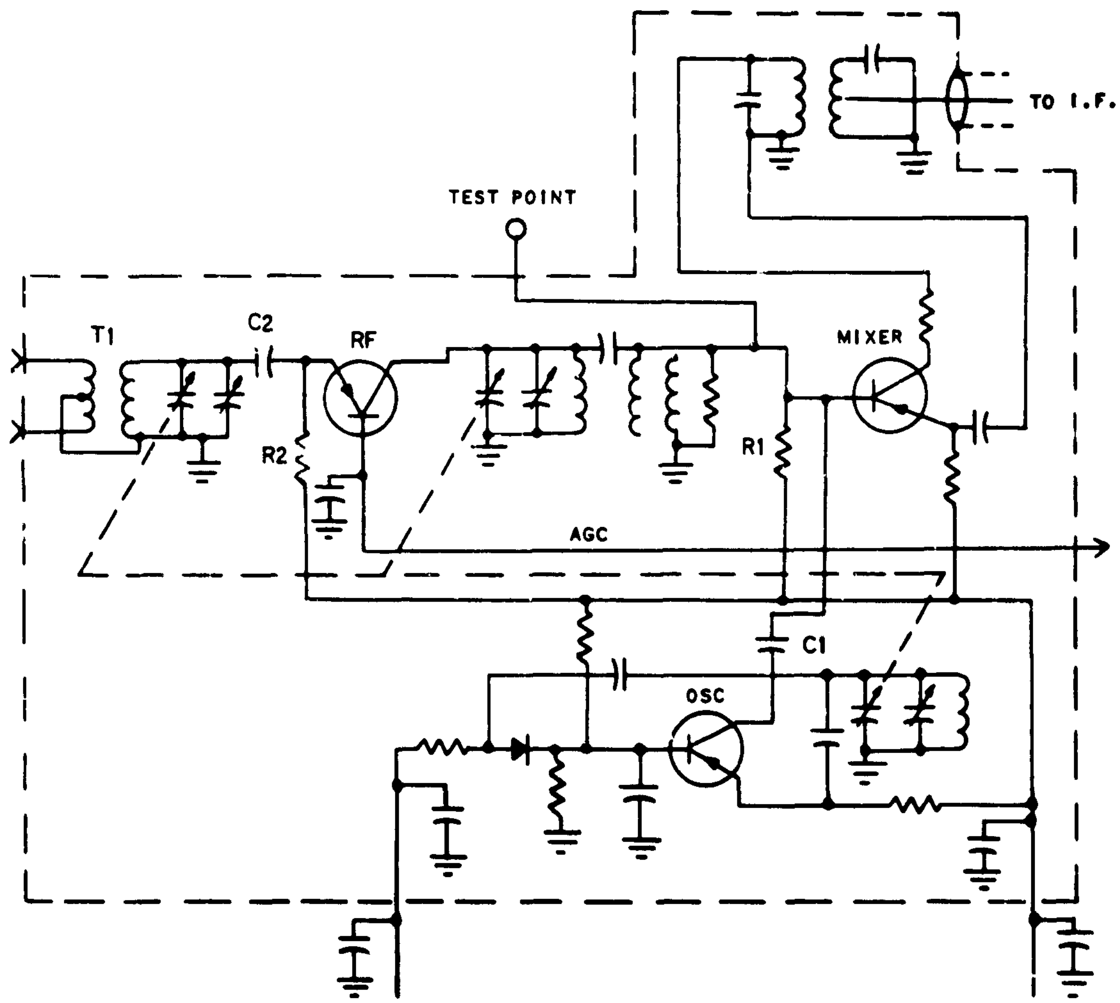


Figure 3-126. Typical Example of Poorly Designed VHF Receiver RF Section

lator signals also appeared on virtually all power and agc circuitry throughout the receiver; and high levels of direct radiation from the receiver and power wiring were found. Also, for high level off-channel signals, the path (R-2, R-1) bypasses the additional tuned circuits in the output of the rf stage, increasing the probability of coupling into the mixer and thus degrading the rejection of spurious signals.

(f) The superheterodyne circuit, employed in receiver applications, is inherently susceptible to interference from certain frequencies other than the frequency to which it is tuned due to spurious responses. In a normal superheterodyne circuit, the desired incoming signal frequency is mixed with the local oscillator frequency in a nonlinear device to produce an intermediate frequency which is accepted by the following receiver amplifier stages. This nonlinear device is termed the first detector, frequency converter, or mixer. A signal of pure sine-wave form contains no harmonics, but when this signal is fed into a nonlinear impedance or nonlinear amplifier, it becomes distorted due to harmonics which occur at integral multiples of the fundamental sine-wave frequency. When two signals are present, such as a signal frequency and local oscillator frequency, an additional heterodyne action occurs in which the sum and difference frequencies are generated between integral multiples of their fundamental frequencies. Therefore, many heterodyne and harmonic frequencies result in the mixing process of the frequency converter stage. This action may also occur in a nonlinear rf amplifier stage due to overloading by a strong signal, improper bias, or improper adjustment of gain. Any of these fundamental, harmonic, or heterodyne frequencies, which occur at the intermediate frequency and reach the mixer stage, may cause interference if their amplitudes are sufficiently high. The sum or difference frequency of the fundamental local oscillator and desired signal frequency is normally selected as the intermediate frequency by means of a resonant circuit in the mixer output. This signal is then further amplified by successive if stages. An illustration of this action is shown in figure 3-127. In general, most cases of interference due to spurious responses of a receiver occur only when a high-amplitude signal is present.

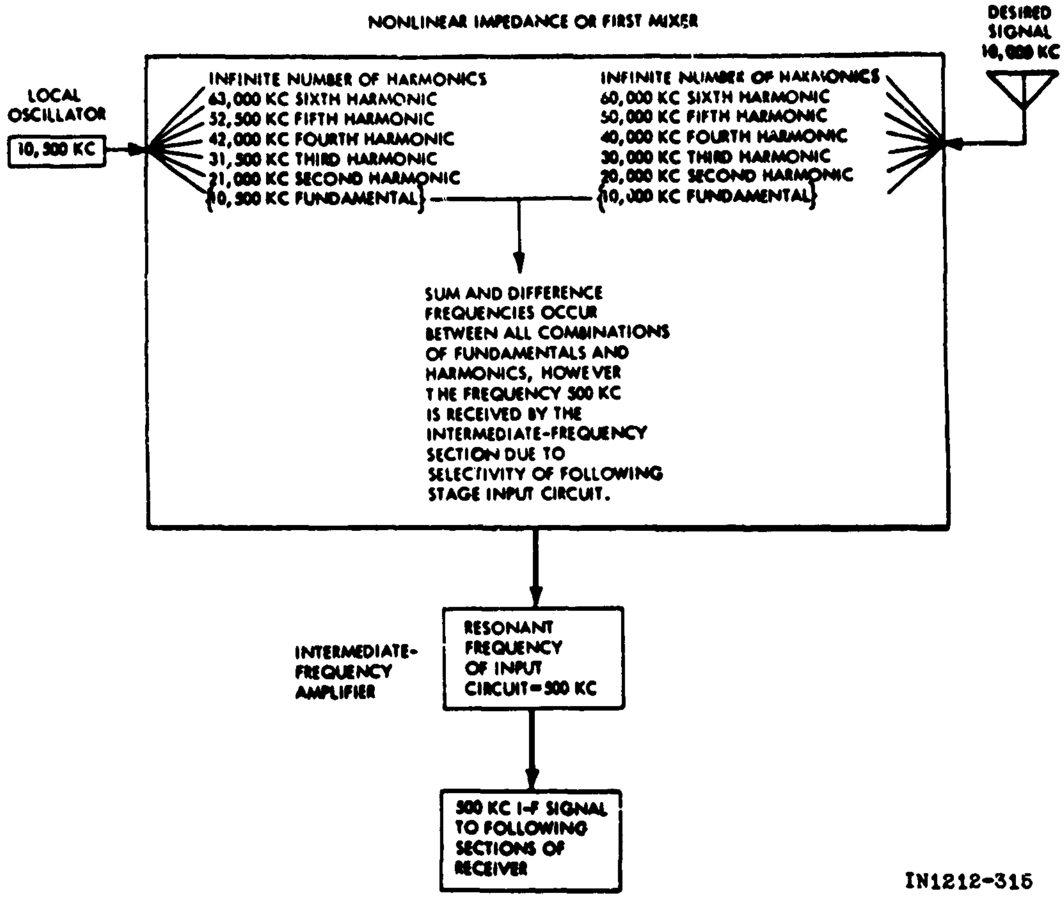


Figure 3-127. Superheterodyne Mixer

(g) Two factors are involved which require an undesired signal to be of high amplitude before it can cause interference. Since most spurious responses are dependent upon harmonic generation and/or heterodyne action within a receiver, the frequency of an interfering signal is usually outside the passband of the rf section of the receiver. Thus, the rf selectivity of a receiver will cause a degree of attenuation to these signals as shown in figure 3-128. It may be noted that attenuation becomes greater as a signal frequency becomes further removed from the center of a passband; however, the attenuation is always some finite value. Therefore any signal, if it is of sufficient amplitude, may pass through the tuned stages of a receiver, but the further removed from the passband, the greater must be its amplitude to pass through. Since higher harmonics of a signal frequency are progressively lower in amplitude, the sum and difference frequencies of higher harmonics are also progressively lower in amplitude. This condition also requires that interfering signals be of high amplitude before they can cause interference due to spurious responses of a receiver.

3-28. Intermodulation

Intermodulation refers to an undesired signal produced in a receiver as a result of the mixing of two or more off-channel signals in a non-linear device. It is one of the most serious forms of interference affecting communications receivers. The importance of preventing this type of interference arises from the fact that reception of a desired signal may be obscured by intermodulation from signals which are removed in frequency from the desired signal.

a. Theory. The mechanism by which the intermodulation occurs is explained through the representation of a non-linear device by a power series. Let the output variable (e) of the device be given as a function

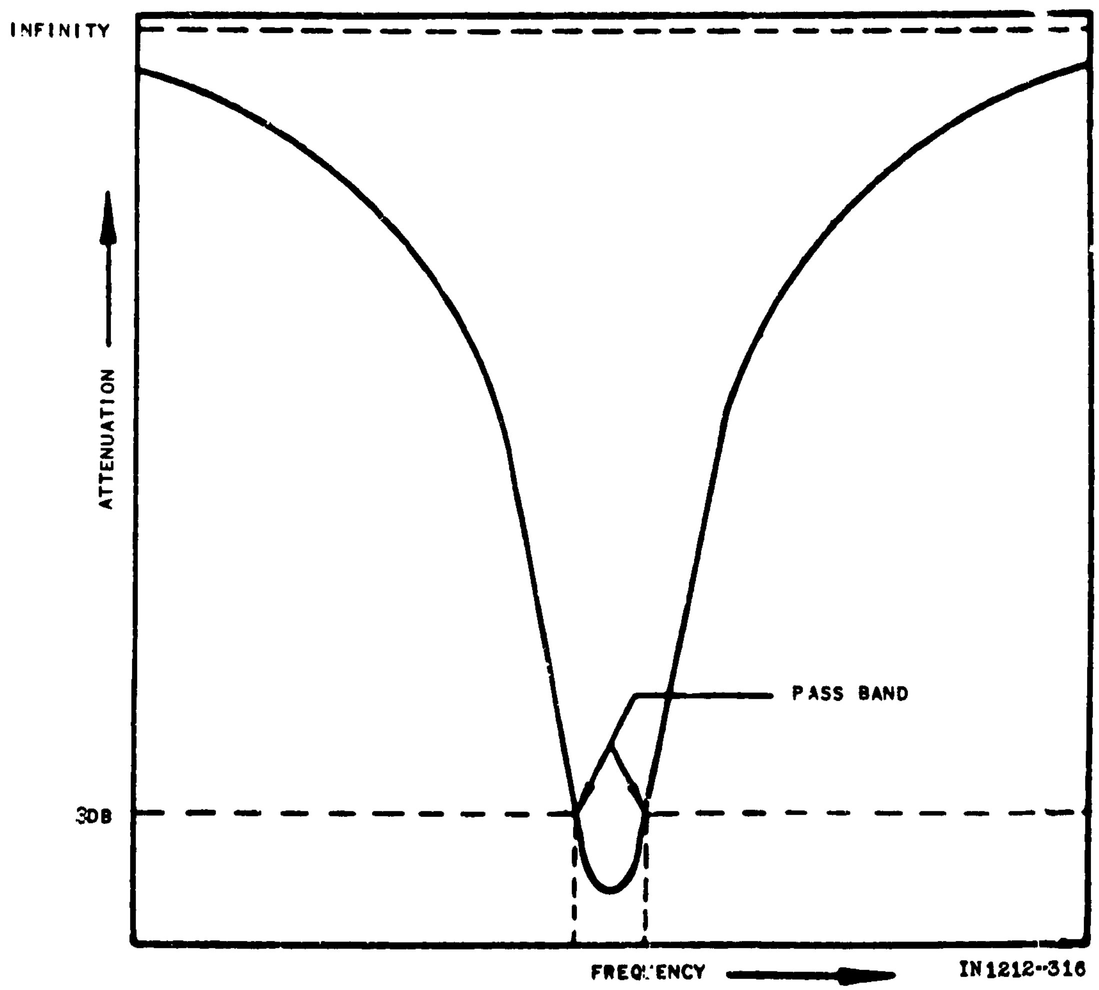


Figure 3-128. Ideal Tuned Circuit Attenuation

of the input variable (e_g):

$$e = a_o + a_1 e_g + a_2 e_g^2 + a_3 e_g^3 + \ldots\ldots. \qquad (3\text{-}56)$$

Now, if the input variable is represented as a sum of two undesired signals, both unmodulated (for simplicity):

$$e_g = e_a \cos 2\pi f_a t + e_b \cos 2\pi f_b t \qquad (3\text{-}57)$$

Upon substituting equation 3-57 into equation 3-56, the following products, involving combinations of the two fundamental frequencies, can be obtained:

$$a_2 e_a e_b \cos(w_a + w_b) t \qquad w_a = 2\pi f_a$$

$$a_2 e_a e_b \cos(w_a - w_b) t \qquad w_b = 2\pi f_b$$

$$3/4\ a_3 e_a^2 e_b \left[\cos(2w_a + w_b) t + \cos(2w_a - w_b) t\right]$$

$$3/4\ a_3 e_a e_b^2 \left[\cos(2w_b + w_a) t + \cos(2w_b - w_a) t\right]$$

Any products resulting from the substitution of two or more signals for e_g in equation 3-56 are referred to as "intermodulation products". Signals yielding the frequency combinations (f_a+f_b) and (f_a-f_b) are termed second-order intermodulation products. Third-order products arise from expansion of the term e_g^3; and n:th-order products from expansion of e_g^n. For two off-channel signals, the third-order product frequencies are $(2f_a \pm f_b)$ and $(2f_b \pm f_a)$.

(1) The theoretical discussion above, relative to the generation of intermodulation products, assumes that the interfering signals have reached a non-linear device or circuit element and is probably of interest to the design engineer only to the extent of demonstrating the processes and the various combinations of signals which, if allowed to reach the mixer for example, might

produce intermodulation interference. In the presence of two extremely high-level signals having the proper frequency relationship to produce sum or difference frequencies in the passband of a receiver, or perhaps the if frequency, intermodulation interference will be produced in almost any receiver. Since no tuned circuits provide infinite attenuation, there will always be some level at which signals can reach the mixer stage. Usually, however, signals of such a magnitude would overload the first rf stage, driving it into non-linear operation and mixing would occur in that stage. It is the design engineer's responsibility to achieve the maximum degree of intermodulation rejection by circuit design, choice of tubes, or solid-state devices, and to insure operation within their linear operating range. It is the responsibility of both the designer and those who specify size and weight requirements to insure that minaturization, for example, is not carried to such extremes as to preclude adequate rf preselection, decoupling, etc., required for reduction of intermodulation.

(2) As previously mentioned, the choice of tubes and the range of operating bias is one of the factors affecting intermodulation interference. However, individual tubes may vary in their intermodulation characteristics at a given bias level. A typical example of third-order intermodulation versus bias and gain is shown in figure 3-129. It can be seen from this figure that published tube characteristics may not be too reliable in that appreciable variations from tube to tube can exist. Nevertheless, the operating bias is seen to affect intermodulation characteristics to the extent of 20 to 30 db.

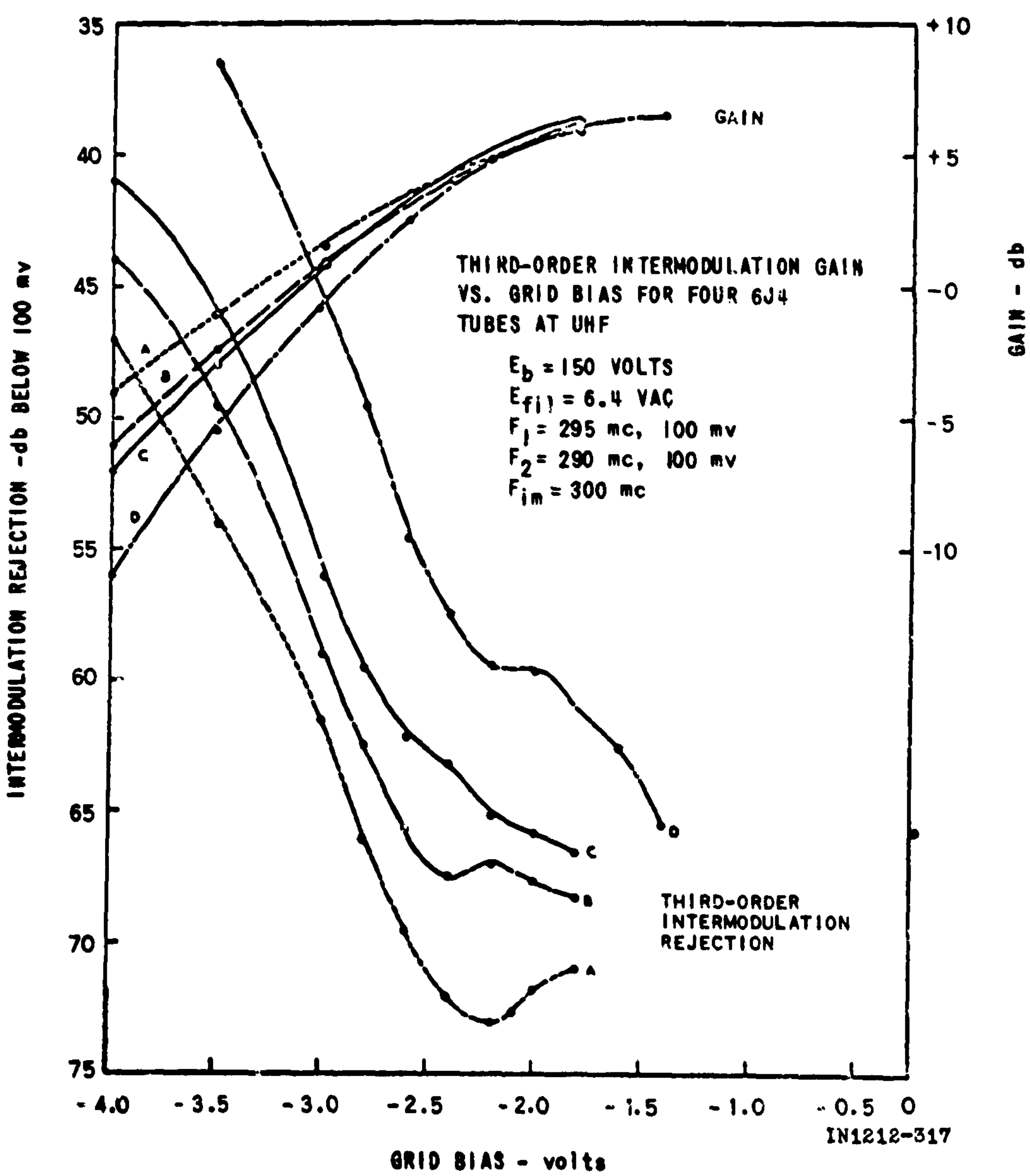


Figure 3-129. UHF Gain and Intermodulation Rejection Characteristics for 6J4 Triodes

3-29. Cross Modulation

Cross modulation is the transfer of the amplitude modulation of an undesired signal to the desired signal. Cross modulation occurs when both the desired signal within the receiver pass-band and the undesired signal are simultaneously present in a non-linear device such as a mixer or overdriven rf stage. As in the case of intermodulation and other spurious responses, the undesired signal must reach the mixer, or an earlier stage must be operating in a non-linear range. The undesired signal may reach the mixer as a result of inadequate isolation in the front-end, allowing coupling of strong signals outside the pass-band to bypass the tuned circuits, or spurious resonances in the tuned circuits. Receivers having no tuned rf stages ahead of the mixer will obviously exhibit poor cross-modulation characteristics. In an fm receiver, the limiter will normally minimize the amplitude modulation impressed upon the desired signal. However, interaction between the two carriers will still introduce fm components and undesired response or distortion of the desired signal.

a. The third-order curvature in the plate-current characteristics in a non-linear stage in the preselector of the receiver is the most common cause of both cross modulation and intermodulation. This effect normally occurs in the first detector (mixer) or, for large input signals, it may occur in the earlier rf stages when the first grid starts drawing current. As the rf stage becomes overdriven, it contributes more heavily to the cross-modulation effect. Assume that the incoming signal voltage is applied to a tube in an am receiver whose transfer characteristic is represented by a power series. Then, when the necessary mathematics are carried through in order to find the desired intermodulation products, it becomes evident that it is necessary to have at least the cubic term present for cross-modulation distortion and that only the amplitude is affected. Hence, cross modulation of the desired signal occurs due to a non-linear characteristic and causes distortion in the amplitude modulation.

b. In general, the cross-modulation characteristics of a receiver vary over its tuning range, particularly if band-changing coils are used. With adequate planning during the initial design stages, some advantage may be gained in optimization of a receiver's freedom from this type of interference. Cross-modulation in an am receiver can be considerably reduced by the application of adequate frequency selectivity circuitry between the receiver and the antenna.

3-30. Other Spurious Responses

In a receiver, spurious signals may manifest themselves as two or more stations being received simultaneously, whistles, squeals, an audio tone, erroneous indications of an instrument, and, where agc circuits are employed, as reduced sensitivity. Additionally, many combinations of input frequencies may result in interference in superheterodyne-type receivers due to the generation of signals at the desired signal frequency or at the intermediate frequency. Any input signal or combination of input signals which results in the generation of the intermediate frequency or the desired signal frequency may cause interference, since the generated signal then progresses through the receiver in the same manner as the desired signal. Various combinations of signals which may form the intermediate frequency or signal frequency are shown in figure 3-130.

a. Broadband Interference. Broadband interference consists of impulse or random rf energy exhibited over a wide frequency spectrum. The response of a receiver to this type of interference is dependent upon its sensitivity (gain) and bandwidth. High levels of broadband interference, however, may also generate spurious responses in the same manner as two or more narrowband signals by the same processes or a combination of these processes. Poor rf selectivity may allow broadband energy components, over a wide frequency range, to reach the mixer. To minimize the receiver response to broadband interference, precautions similar to those required to minimize other types of spurious responses should be undertaken, with

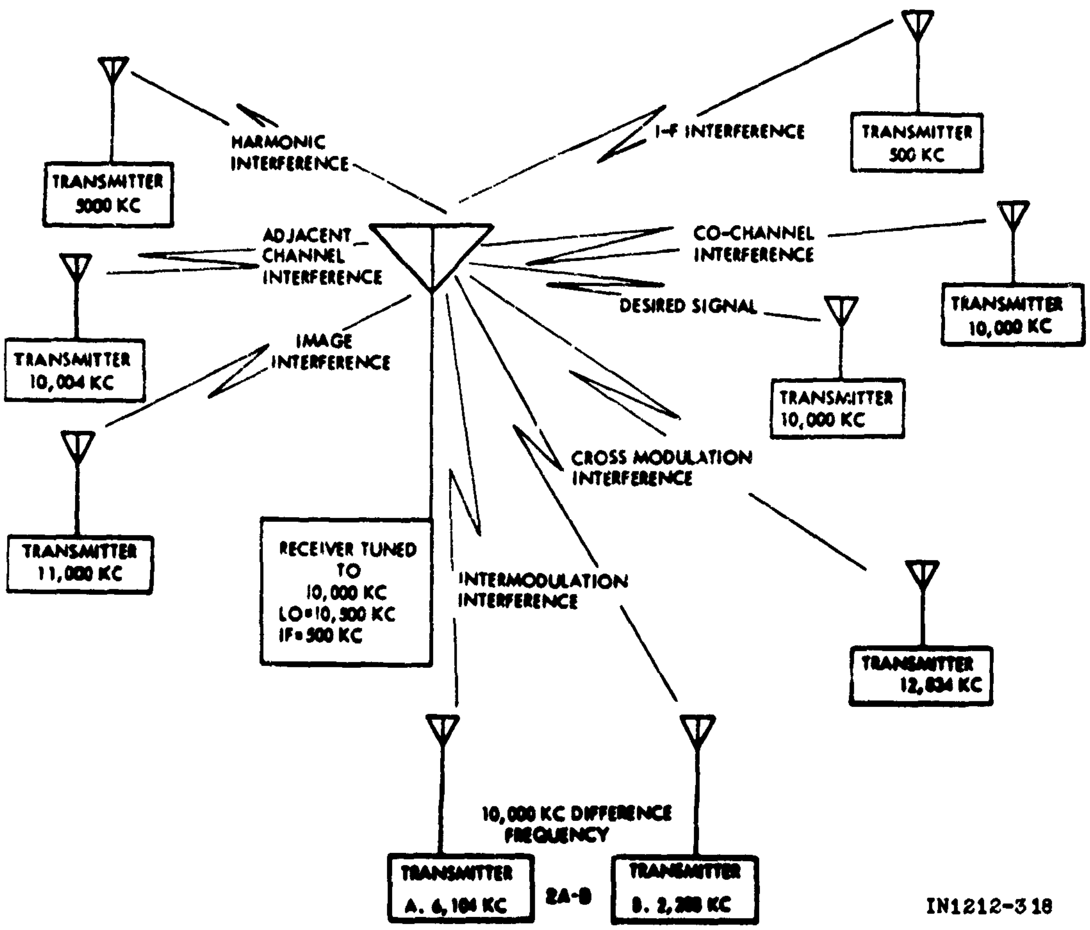


Figure 3-130. Potentiε' Sources of Narrow-Band Interference

the additional consideration of maintaining bandwidth at the minimum required to pass the desired signal intelligence. High Q, tuned circuits with a minimum of spurious resonances will minimize the peak energy reaching the first grid or transistor which could cause spurious responses.

b. Channel. A channel is defined as a band of frequencies of a width sufficient to contain a carrier and/or the minimum necessary sidebands to convey intelligence. The frequency spread of a channel will be dependent upon the type of emission from the transmitter; and the receiver's overall bandpass must be sufficient to accommodate the channel width. For example, in the case of amplitude modulation where the highest modulation frequency is 3.5kc, the channel width would be a minimum of 7kc so as to accommodate the upper and lower sideband frequencies separated by the carrier frequency. Therefore, a receiver designed to receive a signal of this type would require an overall minimum bandpass of 7kc for proper operation as shown in figure 3-131A. In the case of single-sideband (reduced carrier transmission with a maximum modulation frequency of 3.5kc), the channel width would be 3.5kc and a receiver would only be required to have an overall minimum bandpass of 3.5kc for proper operation as shown in figure 3-131B. Frequency-modulated signals have a greater spectrum width, and a receiver must have a sufficient pass-band to accept the sidebands involved. If the spectrum width were 400kc, the required minimum overall bandpass of a receiver for proper operation would be 400kc as shown in figure 3-131C.

c. Linear and Non-Linear Responses. When considering means for analyzing receiver spurious characteristics, it is convenient to evaluate the effects of interference in terms of "small-signal" responses and "large-signal" responses. The mechanism of any interference-reducing scheme (other than blanking) is that:

1) Interfering components of the undesired signal are removed from the bandpass of the susceptible receiver

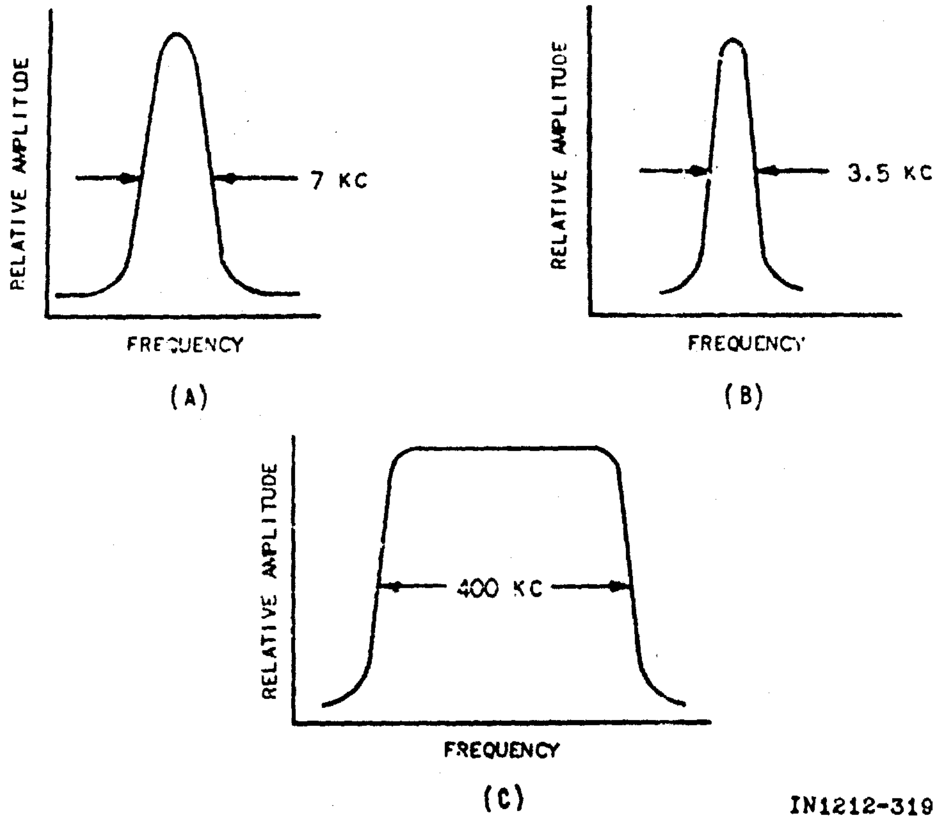


Figure 3-131. Typical Bandpass Curves

2) Non-linear effects, such as overloading and cross modulation, are reduced by the sideband selectivity of the radio-frequency stages

When a signal is small compared to the maximum signal for which a receiver is designed, it is relatively safe to assume linear operation of the receiving device. Signals, which are large compared to the normal signal level for which the receiver is designed, frequently will drive the receiver into its non-linear region. Such non-linear operation gives rise to a large number of spurious interference responses which would not otherwise occur.

d. Self-Spurious Responses. A self-spurious response may occur in two ways:

1) Two locally-generated frequencies, or their harmonics, mix in a non-linear stage to produce either the receiver tuned frequency or the intermediate frequency
2) An oscillator harmonic coinciding with the receiver tuned frequency or an intermediate frequency

Thus, in effect, the receiver interferes with itself. This type of interference occurs in multiconversion receivers where the basic oscillator frequency is multiplied to provide the desired mixing frequency.

(1) Harmonics of the local oscillator often provide the means for spurious responses at frequencies well above the tuned frequency. These responses may occur when, due to poor front-end design or lack of consideration for response of the tuned circuits far out of the pass-band or tuning range, signals pass through, or are coupled around, the rf-tuned circuits and mix with a harmonic of the local oscillator to produce a difference frequency equal to the if frequency. In some cases, a low-pass filter in the output of the oscillator may be necessary. Lower oscillator power often reduces the harmonics also. Oscillator harmonics should be checked in the design or breadboard stage and measures taken to reduce them. While normally lower in level, oscillator harmonics, particularly in the vhf and uhf range, become more difficult to contain by shielding and the decoupling networks designed for the oscillator fundamental may be ineffective at the harmonic frequencies. This form of interference will only occur when an input is present, but can appear as co-channel interference to low-level desired signals or garbled modulation of the desired signal. Careful shielding and decoupling, as in the case of many other spurious responses, is necessary to preclude this type of interference.

3-31. Image Response

In a superheterodyne receiver there are two signal frequencies, one higher and one lower than the local oscillator frequency by an amount equal to the intermediate frequency. They may combine with the local oscillator frequency in the first mixer to produce an if signal as shown in figure 3-132. Either signal frequency may be selected by the rf circuits preceding the first mixer. The selected signal is termed the desired signal, and the other signal is termed the image-frequency signal. Without any pre-selection, the receiver will respond equally to signals both above and below the local oscillator frequency. These responses will differ from the local oscillator frequency by the if frequency. This "image" interference is located at frequencies equal to the tuned frequency plus or minus twice the intermediate frequency:

$$f_u = (f_s \pm 2f_{if}) \tag{3-58}$$

where:

f_u = interfering frequency

f_s = desired frequency

f_{if} = intermediate frequency

The appropriate sign is chosen depending on whether the local oscillator frequency is above or below the desired signal.

a. The frequency response of a superheterodyne whose rf input circui is tuned to a frequency, f_s, of 1,000 kcps, is indicated in figure 3-133. First, assume that the oscillator frequency, f_o, is 1175 kcps and that the intermediate frequency, f_{if}, is 175 kcps. If there should be an incoming transmission at the frequency f_{im} = 1,350 kcps, figure 3-133A, it will cause a difference frequency of 1,350 - 1,175, or 175 kcps in the converter stage and will result in a receiver response. Such an undesir signal is known as in image signal or image, and the frequency $f_{im} = f_s$ $2f_{if}$ is known as the image frequency. The image rejection ratio of a r ceiver is the ratio of the input required at the frequency f_{im} to t

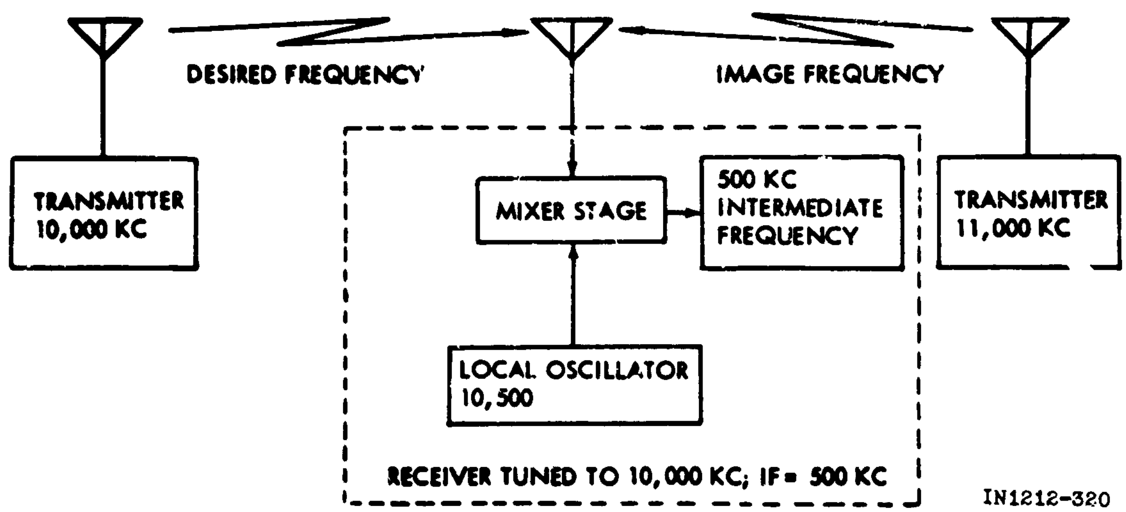
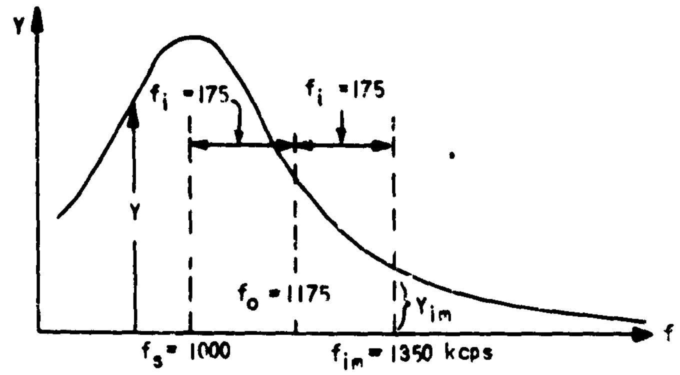


Figure 3-132. Image-Frequency Interference

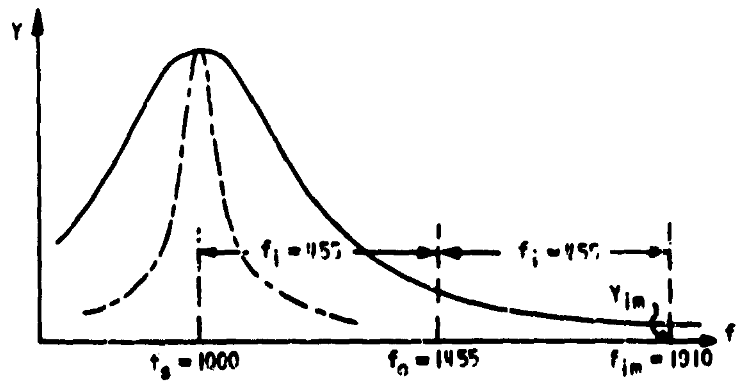
required at the frequency f_s to produce output signals of the same strength. The image rejection ratio must be high for image-channel interference, or for image interference, to be negligible.

b. To secure a high image rejection ratio, the intermediate frequency should be made as large as other considerations permit. An intermediate frequency of 455 kcps is used frequently in modern broadcast receivers. As seen in figure 3-133B, the admittance, Y, of the receiver at the image frequency of 1,910 kcps is extremely small. However, an increase in intermediate frequency generally reduces the selectivity and the stability of the receiver.

c. Image response is minimized by increasing the selectivity of the rf stages of the receiver by additional tuned circuits or additional rf stages. Careful design of the tuned circuits is necessary as in the case of reduction of any other type of spurious response. If the rf input response curve is made sharper, as indicated by the dot-dash curve in



A. IF = 175 KC



B. IF = 455 KC IN1212-321

Figure 3-133. Image Frequency Response in a Converter figure 3-133B, the image rejection ratio is increased. However, the response curve of the input circuits cannot be made too narrow or side-band clipping may then be encountered. The tubes in a preamplifier, in themselves, do not contribute to the image-suppression qualities, so in many instances, additional tuned circuits may be added without necessitating rf amplification.

d. In the higher frequency ranges, where good rf selectivity becomes more difficult to achieve, the choice of a higher if frequency (thereby removing the image frequencies farther from the center frequency) is desirable and often essential to minimize image response. Double conversion often becomes necessary to obtain an image frequency sufficiently far removed from the center frequency. In this case, the first if frequency is made as high as practical, with the second if at a lower frequency to achieve the necessary gain. The rf-tuned stages, however, must still be carefully designed and checked for possible spurious resonances which might occur even at a far-removed image frequency. Performing rf selectivity measurements over only a very limited range, possibly down to the 60 db points above and below the center frequency, is a common error. RF selectivity of a front-end design should be checked upward to at least 2×if above the 5th harmonic of the local oscillator (lo), and down to the lowest frequency capable of mixing with the basic lo frequency, to insure minimizing image and other spurious responses.

e. Another form of image response is produced when an undesired signal reaches the mixer and mixes with a harmonic of the local oscillator. A signal at a frequency $\pm f_{if}$ removed from a harmonic of the local oscillator can produce an if signal out of the mixer and hence a spurious response. At first glance, this would not appear to be a serious consideration, since it concerns a signal far removed in frequency from the center or tuned frequency. However, such responses have been recorded up to the seventh harmonic of the local oscillator mixing frequency and at levels as low as 50 millivolts at the antenna terminals of the receiver. This is attributed to two basic causes; excessively high oscillator harmonics and poor front-end design allowing signals, far out of the theoretical pass-band, to reach the mixer. Such problems may be minimized by reduction of local oscillator power, thus improving the waveform; or utilizing a low-pass filter in the output of the local oscillator. Again, spurious resonances, or coupling around the rf and mixer tuned circuits, are a primary cause and should be carefully checked and corrected in early design or breadboarding of the receiver front end. Electrostatic shields between windings of rf transformers, and other precautions to minimize capacitive coupling, aid in minimizing these responses. Minimizing lead lengths throughout tuned circuits and avoiding common primary and secondary return leads, however short (which become increasingly inductive at frequencies several times the center frequency), also help reduce this problem. Where spurious resonances above the tuning range cannot be avoided due to inherent component limitations, then low-pass filters should be employed, preferably ahead of the first rf stage. This practice also increases intermodulation rejection in the higher frequency ranges.

3-32. Oscillators and Mixers

a. <u>Combination of Different Harmonics of the Signal and Oscillator.</u> Combination of the interference signal and local oscillator harmonics of different integers, of the form;

$$\pm m f_{het} \pm n f_u = f_{if} \tag{3-59}$$

where f_{het} = local oscillator frequency

f_u = interfering frequency

f_{if} = intermediate frequency

results in interference responses which are spread over the tuning range in an irregular manner. This interference is due to the fact that the frequency changer, or mixer, does not have an ideal grid-voltage/plate-current characteristic, and thus produces harmonic outputs.

b. Oscillator Multiplication. A common design practice, especially with crystal-controlled receivers, is to multiply the crystal frequency to obtain the desired oscillator frequency. Consider the contribution to the interference problem by multiplication. Assume that an oscillator frequency of 80 megacycles is desired, and that a fundamental crystal is to be used. Two stages of multiplication could be used, e.g., a doubler followed by a doubler. Thus, the crystal frequency would be 20 megacycles. Unless elaborate measures are taken, such as using a double-tuned amplifier stage between doublers and a double-tuned amplifier stage between the second doubler and the mixer, the output will contain signals of 20 mc and 40 mc as well as 80 mc. This provides two additional frequencies at the mixer with which spurious responses may be generated. A way to decrease the number of possible interfering frequencies is to decrease the multiplying factor (increase the crystal frequency) to as low a value as is practically and economically possible. Other methods of reducing the interference phenomena caused by multiplication are:

1) Shield all oscillator stages
2) Decouple all plate and filament circuits
3) Provide multiple tuned circuits or bandpass filters so that the undesired harmonics of the crystal frequency are not injected into the mixer

c. Oscillator Harmonics. Harmonics of the local oscillator frequency which may be generated in the mixer tube can combine with signals leaking through the rf stages of a receiver in such a manner that their sum or difference frequency is equal to the if of the receiver. These difference frequencies represent spurious responses, and their occurrence may be predicted theoretically. It is sometimes helpful to have a set of quick reference curves to identify spurious responses. The combinations possible from all different types of receiver mixing are derived below. Theoretical spurious response curves for commonly encountered if frequencies of 455 kc, 10.7 mc, and 30 mc are used as examples. Refer to figures 3-134 through 3-142 inclusive.

(1) Derivation: Let f_o be the local oscillator frequency, f_i be the if, f_c be the frequency of the interfering signal, and f_d be the dial-setting of the receiver. The receiver will respond to any signal that reaches the intermediate frequency stage. That is if:

$$\pm nf_o \pm f_c = f_i \qquad (3\text{-}60)$$

a response will occur. Here n is an integer, and nf_o is a harmonic of the local oscillator frequency.

(a) Receiver if's are designed to respond to either sum or difference frequencies. Therefore, three different types of response are possible.

Type A: $$f_o = f_i - f_d \qquad (3\text{-}61)$$

Type B: $$f_o = f_d - f_i \qquad (3\text{-}62)$$

Type C: $$f_o = f_d + f_i \qquad (3\text{-}63)$$

(b) If harmonics of the incoming signal are generated, it is possible to produce additional responses. These can be included in the derivation by adding an integer multiplier on f_c in the first equation above. However, this feature will be omitted from this derivation since, at the present time, these harmonics do not seem to be a serious source of spurious responses.

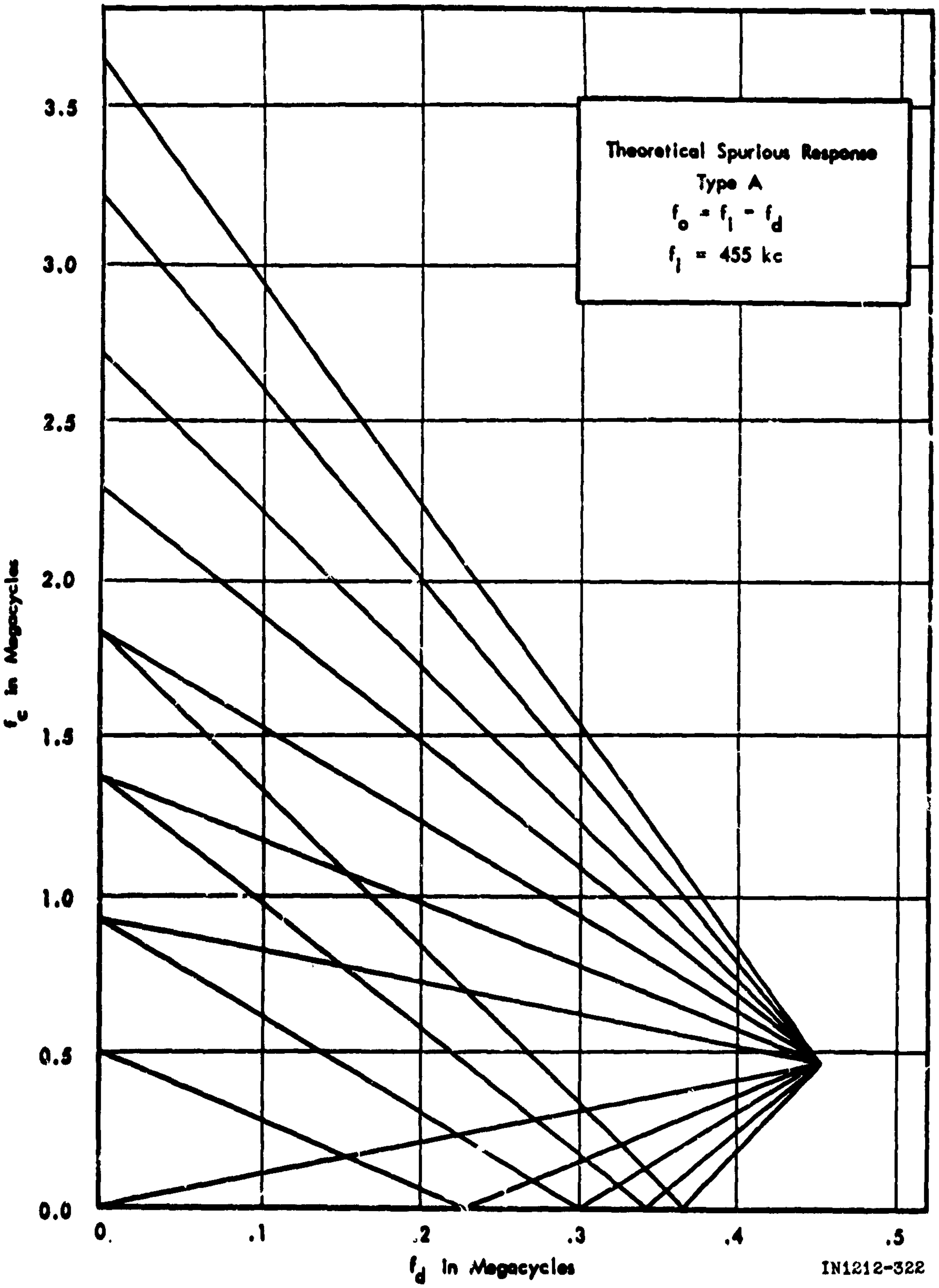


Figure 3-134. Theoretical Spurious Response, Type A, IF = 455 KC

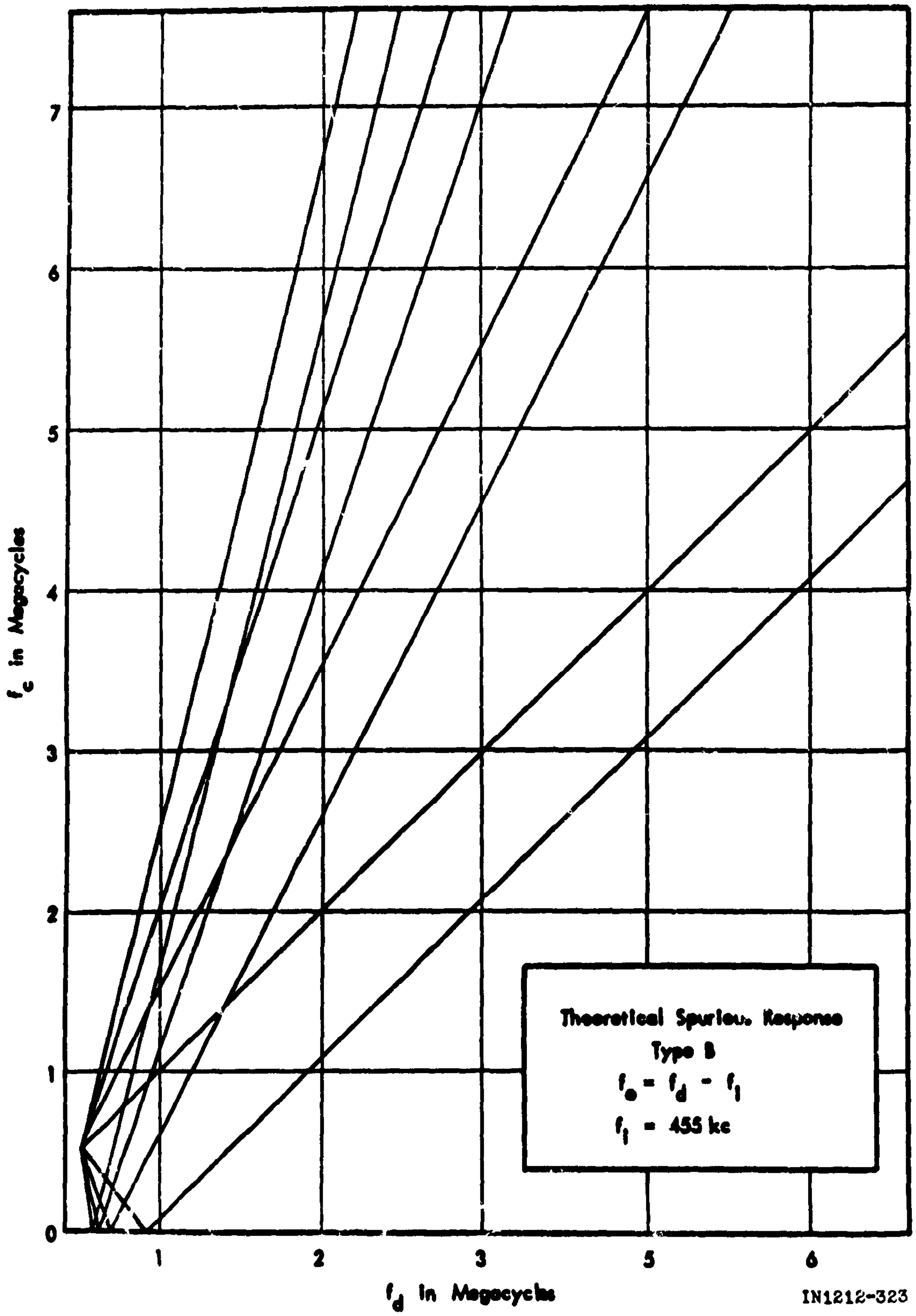


Figure 3-135. Theoretical Spurious Response, Type B, IF = 455 KC

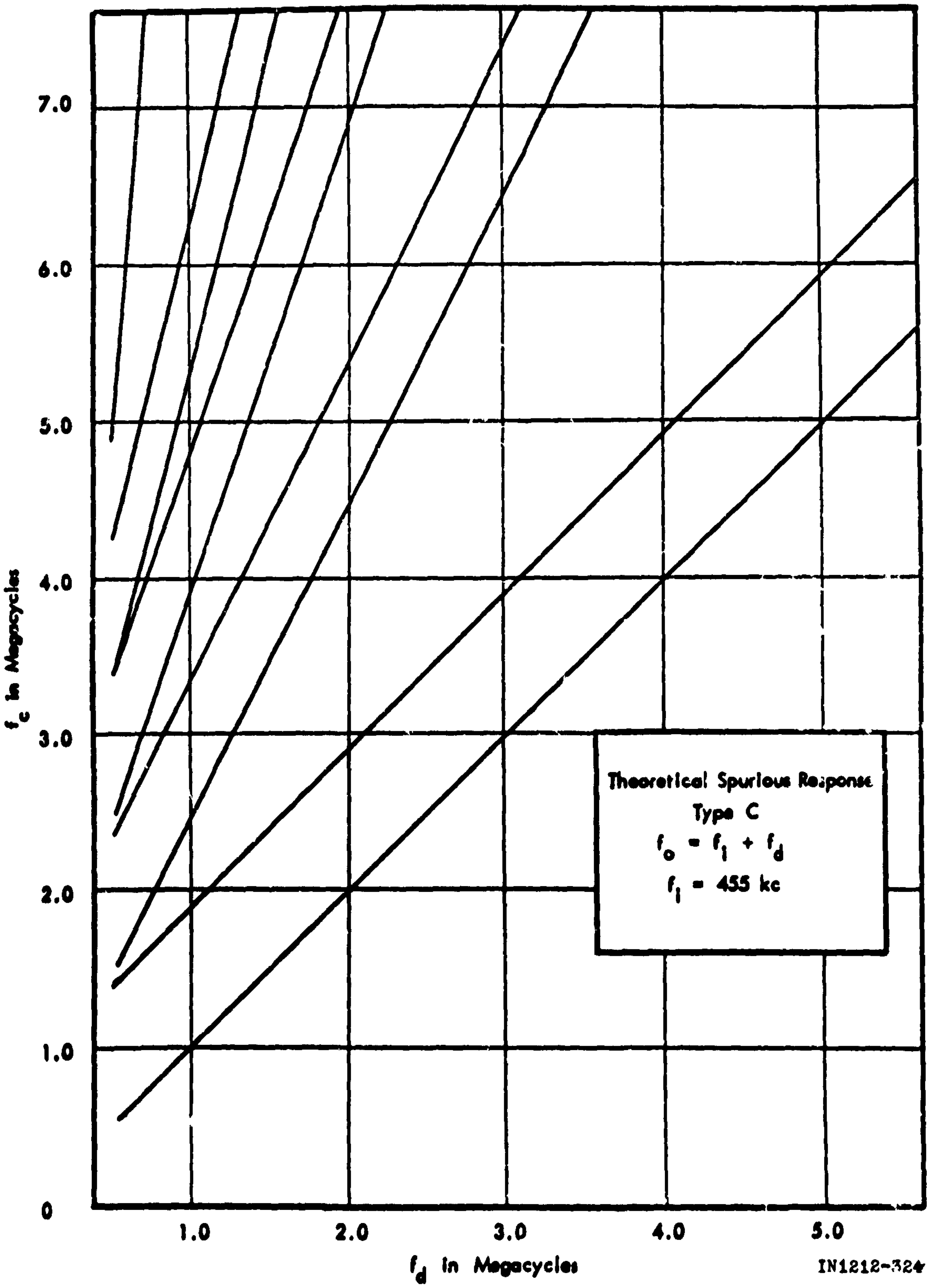


Figure 3-136. Theoretical Spurious Response, Type C, IF = 455 KC

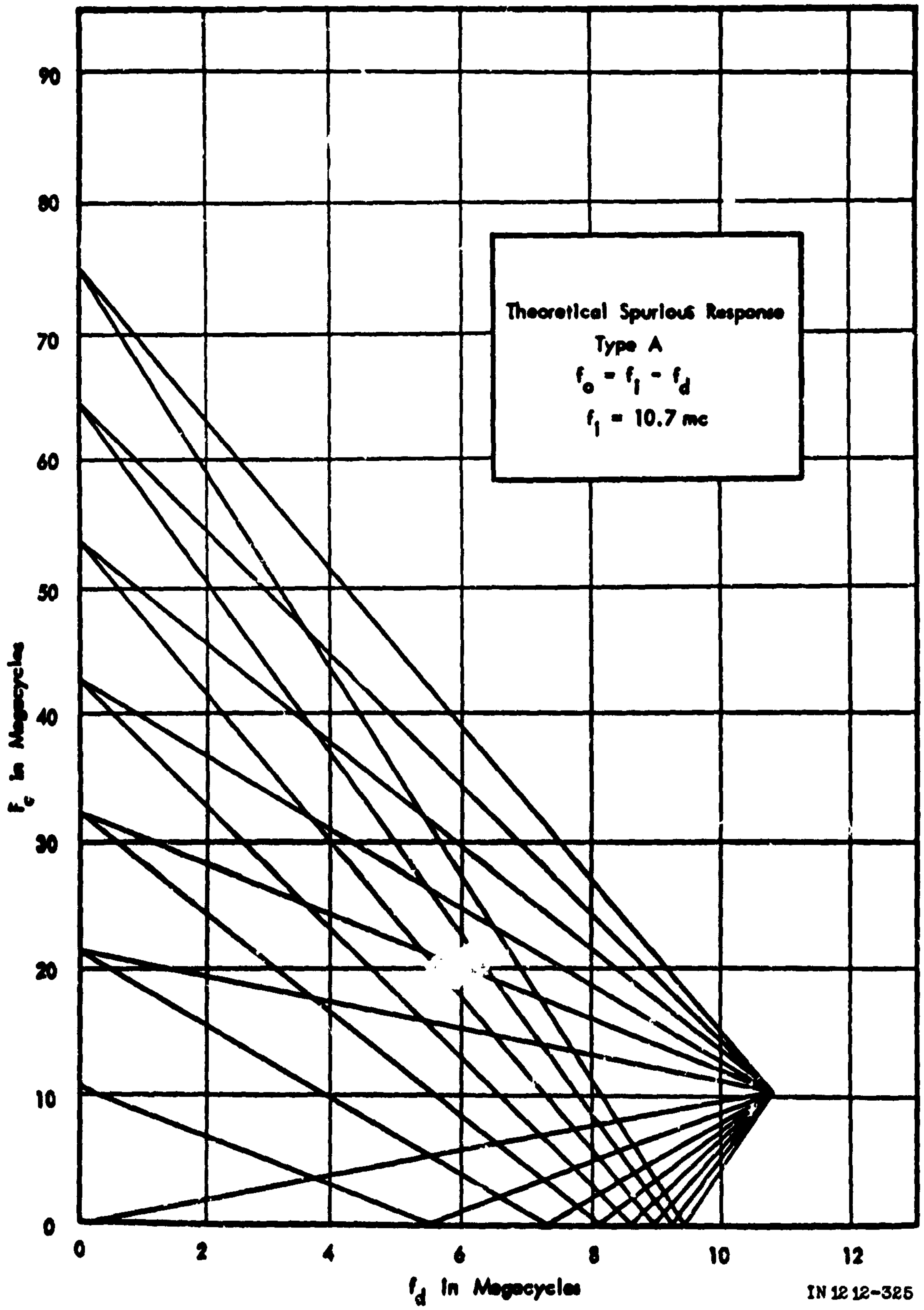


Figure 3-137. Theoretical Spurious Response, Type A, IF = 10.7 MC

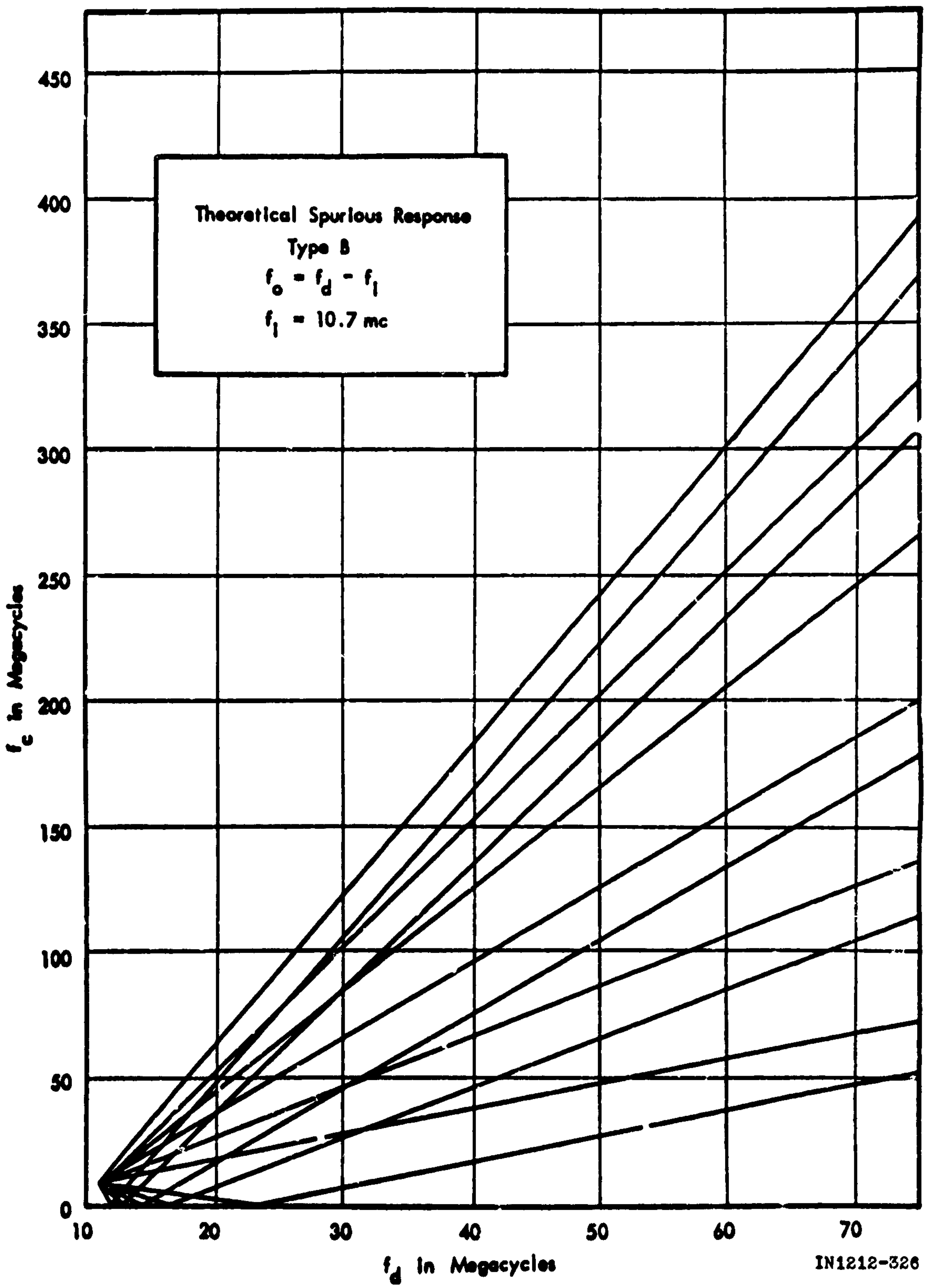


Figure 3-138. Theoretical Spurious Response, Type B, IF = 10.7 MC

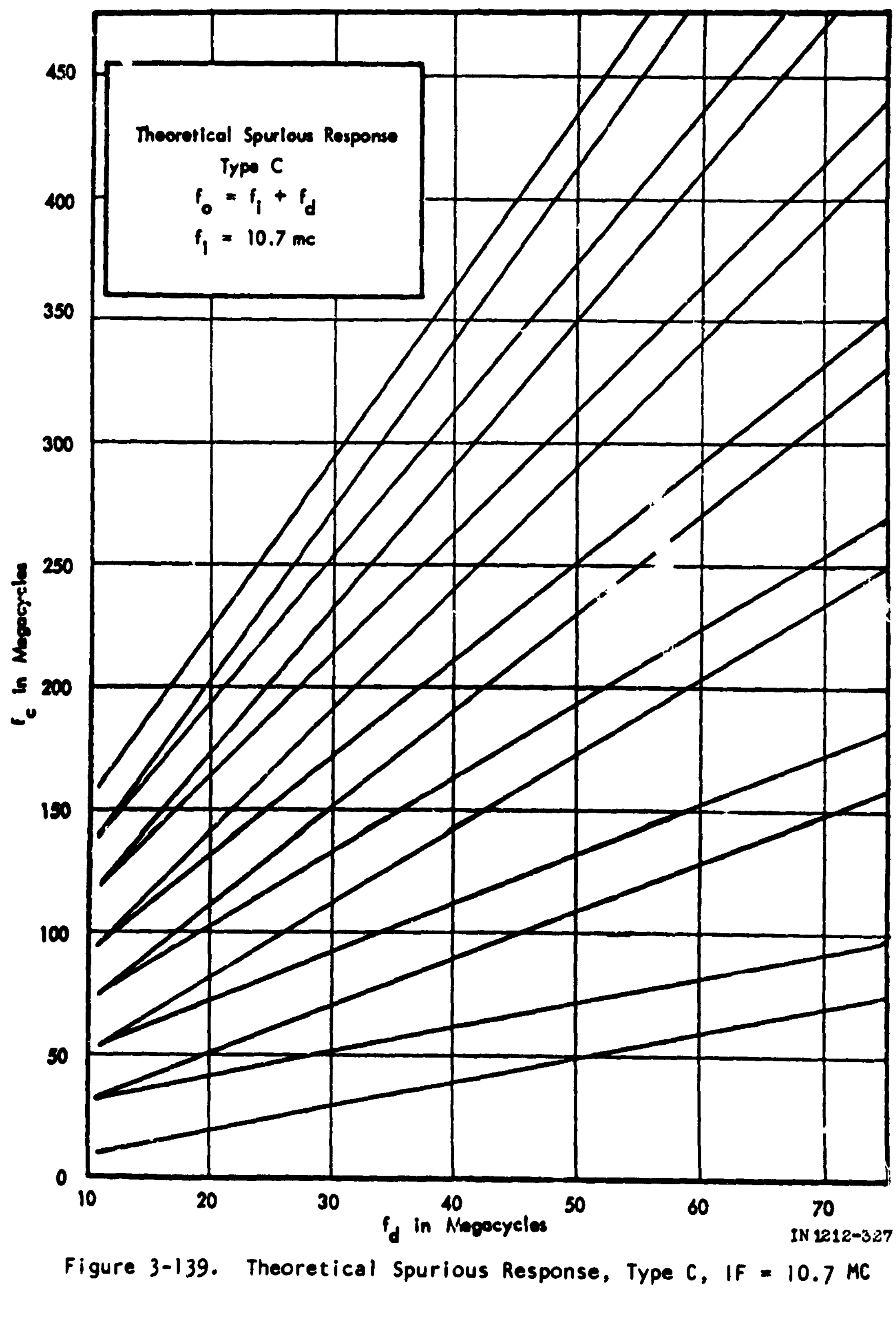


Figure 3-139. Theoretical Spurious Response, Type C, IF = 10.7 MC

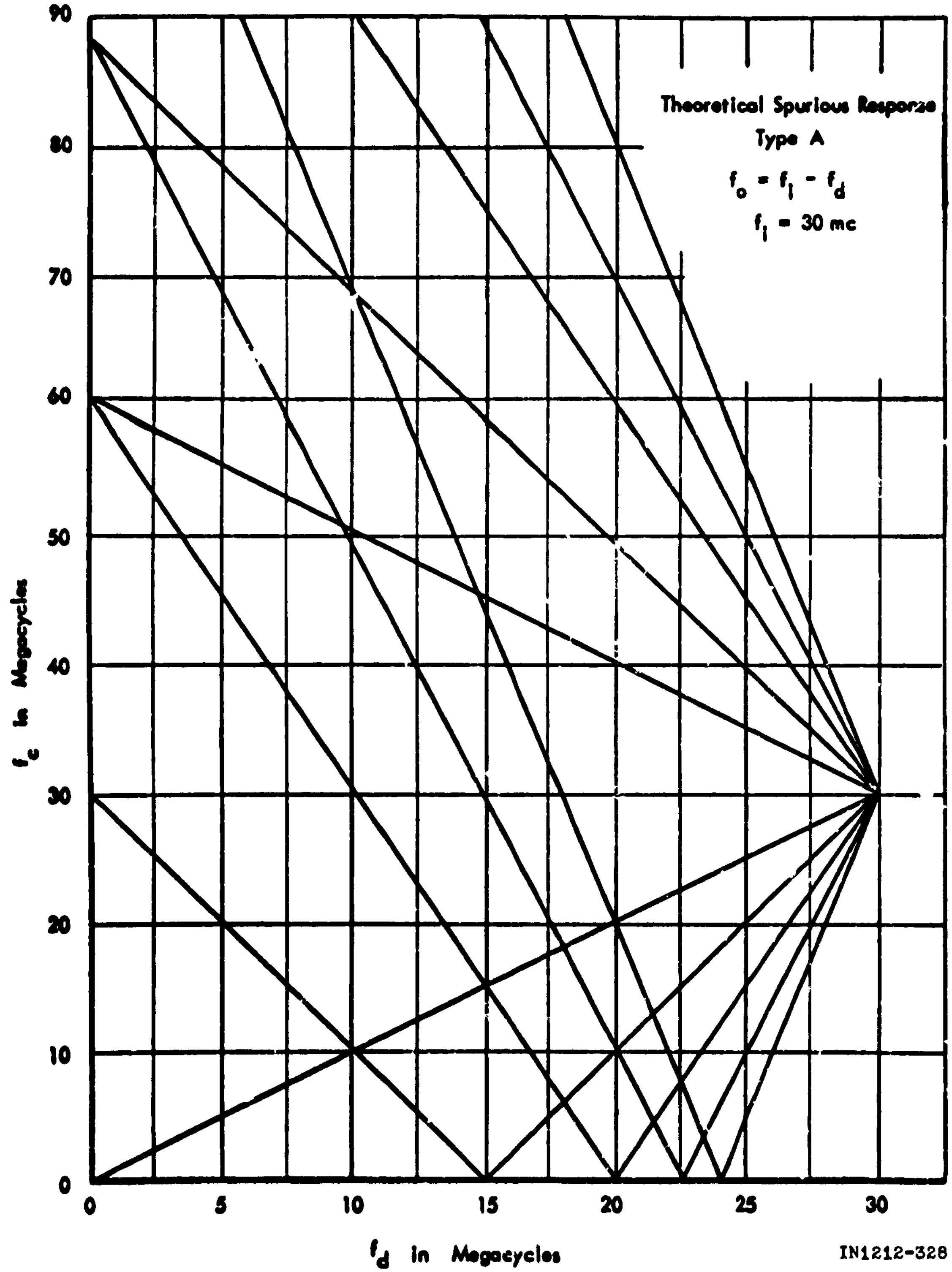


Figure 3-140. Theoretical Spurious Response, Type A, IF = 30 MC

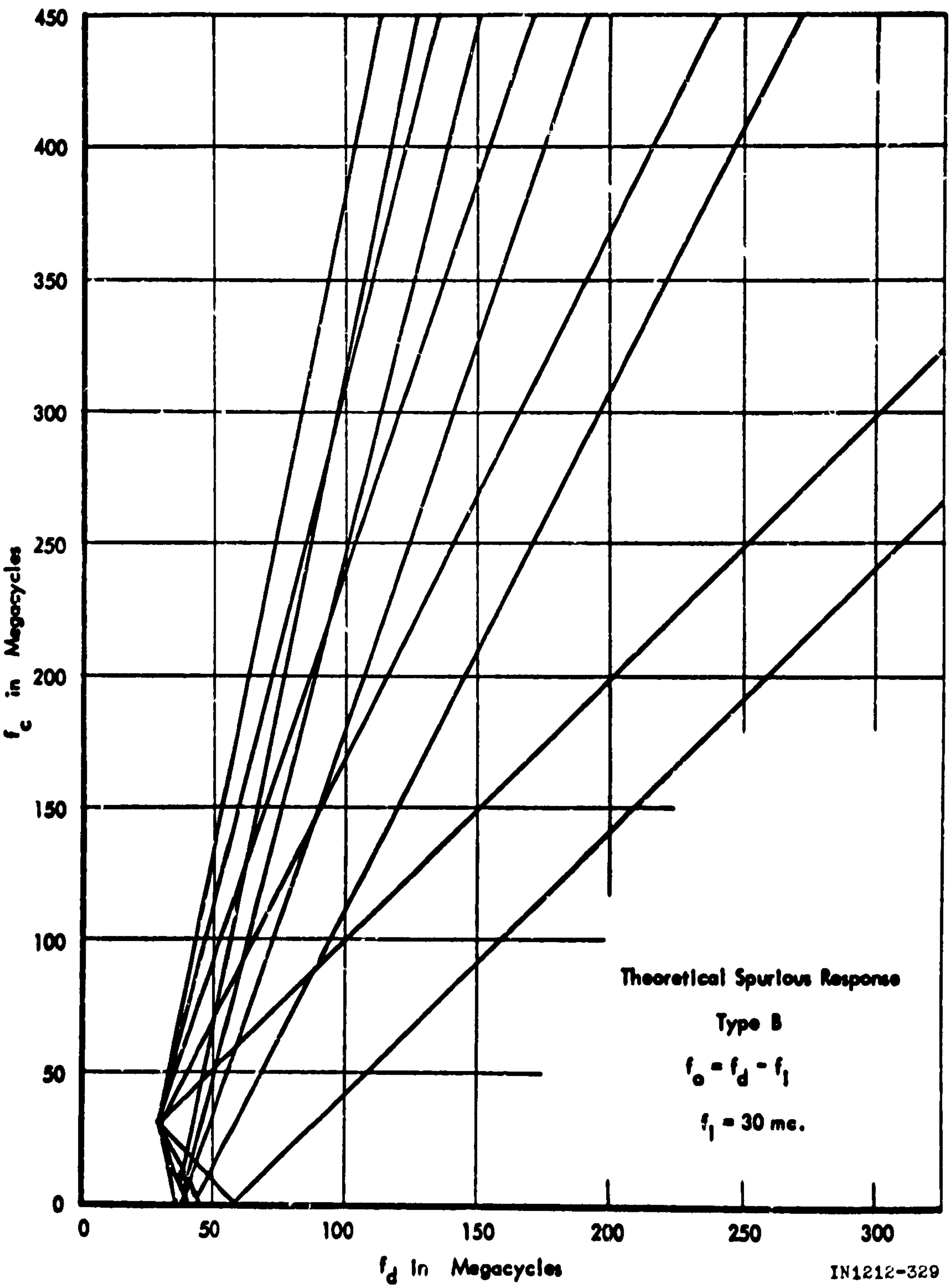


Figure 3-141. Theoretical Spurious Response, Type B, IF = 30 MC

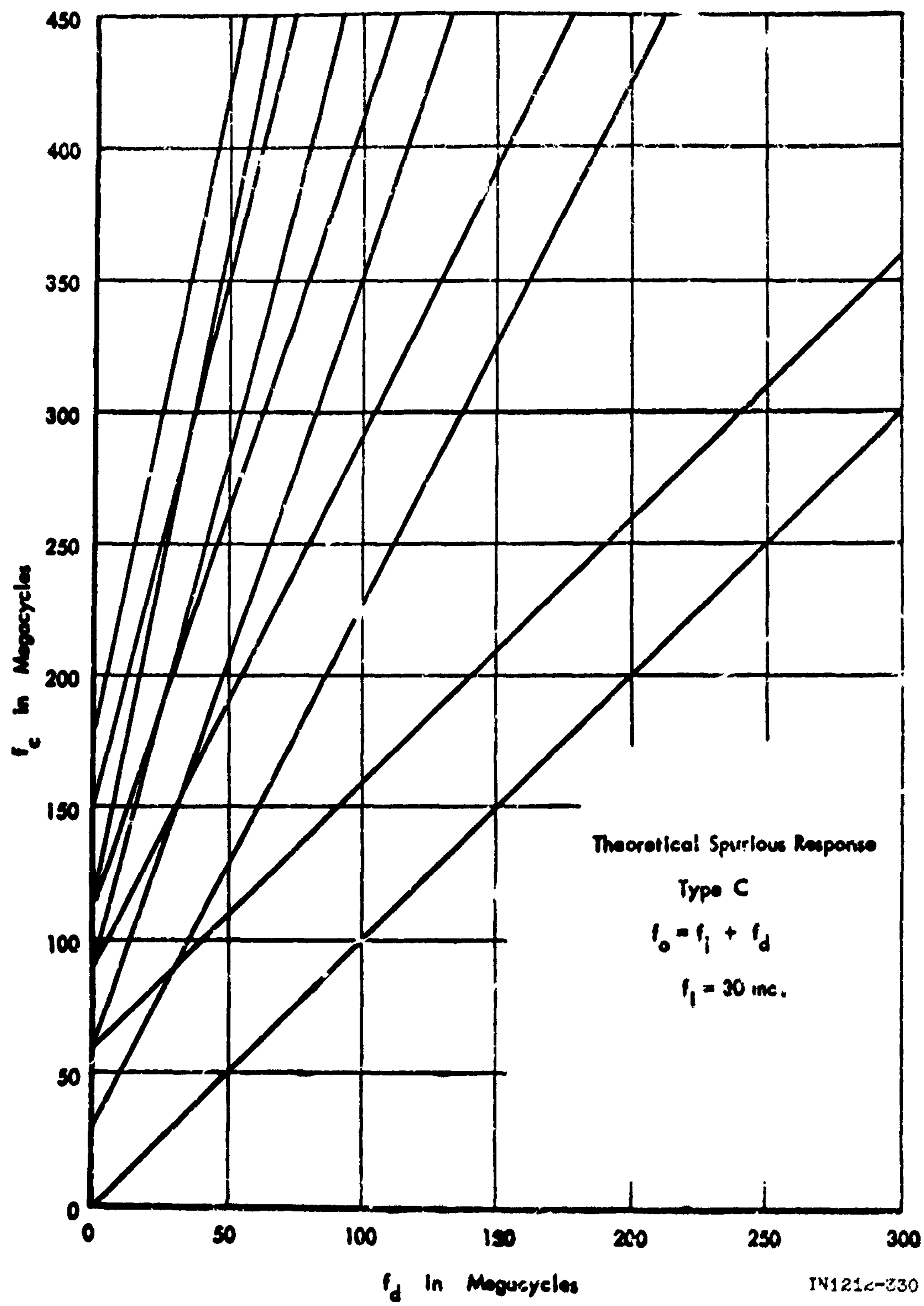


Figure 3-142. Theoretical Spurious Response, Type C, IF = 30 MC

(c) For each type of operation, the possible spurious response frequencies are obtained by substituting the equation representing the type of operation into the first equation (3-60) and solving for f_d. In each case, four equations are obtained of which one or more represent impossible conditions. The usable resulting equations are:

For Type A:

$$f_d = \frac{n-1}{n} f_i + \frac{f_c}{n} \quad (3\text{-}64)$$

$$f_d = \frac{n-1}{n} f_i - \frac{f_c}{n} \quad (3\text{-}65)$$

$$f_d = \frac{n+1}{n} f_i - \frac{f_c}{n} \quad (3\text{-}66)$$

For Type B:

$$f_d = \frac{n+1}{n} f_i - \frac{f_c}{n} \quad (3\text{-}67)$$

$$f_d = \frac{n+1}{n} f_i + \frac{f_c}{n} \quad (3\text{-}68)$$

$$f_d = \frac{n-1}{n} f_i + \frac{f_c}{n} \quad (3\text{-}69)$$

For Type C:

$$f_d = \frac{1-n}{n} f_i + \frac{f_c}{n} \quad (3\text{-}70)$$

$$f_d = \frac{f_c}{n} - \frac{n+1}{n} f_i \quad (3\text{-}71)$$

d. Mixers. The process of combining two unrelated signals to obtain a third signal at the sum or difference frequency has been employed in electronic devices for many years. The circuits which accomplish the combining process generally produce unwanted products as well as the required signal. The unwanted products are generated by nonlinearities within, or prior to, the mixing circuit, and are classified as products of nonlinear mixing.

(1) Linear mixing may be defined as that form of mixing which produces only the sum and difference frequency of the two input fundamental frequencies. It produces no harmonics or other

mixing products and is the desired, but thus far unattained, type of mixing. The analog multiplier more nearly approaches a truly linear mixer than any other known device, but only at very low frequencies of operation.

(2) Nonlinear mixing is the form of mixing presently used for heterodyning two frequencies to obtain the sum or difference frequencies. The resultant by-products of such mixing are the sources of many spurious responses. Considerable effort has been directed toward the problems of nonlinear mixing for many years. A major problem in establishing a mixer's conformity to current theory appears to stem from the difficulty in obtaining accurate experimental data. Many factors contribute to measurement error in attempting to isolate the action of mixer nonlinearity. Depending upon relative rejection, the injected harmonics of either the local oscillator or the input signal can have a highly detrimental effect on the mixer measurement accuracy. The same problems which are encountered in a practical receiver are generally encountered in attempting to perform accurate measurements.

(3) Innumerable problems can be generated within, or associated with, the mixer, most of which are a result of the very mixing process itself. It has been found that harmonics above the 50th can cause a response. On this basis, it would appear impossible to eliminate this phenomenon since non-linearity of a mixer is needed to provide the frequency conversion and it is this very non-linearity which generates the harmonics. However, several precautions may be taken in the mixer design which will minimize the spurious response problems considerably.

(a) Low injection voltages. A very low injection voltage from the oscillator should be used. This will reduce the power of the higher oscillator harmonics, thus reducing spurious responses due to oscillator harmonics. In one receiver, it was found that the number of spurious responses could be lowered from 101 to 28 simply by reducing the injection voltage from 510 millivolts to 42 millivolts

(b) Isolation. Isolation should be used between the rf signal and the oscillator signal. This can be done in several ways; one is through the use of a double triode mixer; a second is through the use of a multigrid mixer; and the third is through the use of high-impedance coupling from the oscillator to the signal grid of the mixer

(c) Balanced mixers. A balanced mixer can be used. This will cancel the even harmonics of the applied rf signal which tend to be generated in the mixer. The possible number of spurious responses will thereby be reduced by one-half. However, it should be noted that, in this case, isolation between the rf signal and oscillator signal would best be accomplished by high impedance coupling

(d) Low-pass filter. Low-pass filters can be placed in the oscillator output to the mixer

3-33. Adjacent Channel Interference

Adjacent channel interference arises from the interaction of an adjacent channel carrier and/or its sidebands with a desired carrier and/or its sidebands. This type of interference may result from inadequate rf selectivity. When this occurs, the adjacent channel signal may be considered as arriving at the detector of the am receiver, or the limiter-discriminator of the fm receiver, without prior distortion or the formation of intermodulation products. In an am receiver, analysis can begin with the summation of the two signals in an am detector such as the square law detector, the product detector, the synchronous detector or the perfect envelope detector; or, for an fm receiver, a perfect limiter and simple discriminator, balanced linear discriminator or a homodyne detector, etc.

a. In amplitude-modulated systems, the adjacent channel response is characterized by the audible "carrier beat", by "monkey chatter" and by "masking" of the desired signal. The unintelligible interference in the audio output known as "monkey chatter" is a result of the inversion of the speech modulation of the undesired signal. This inverted speech is produced on the desired channel by the difference frequencies between the desired carrier and the nearer adjacent channel sideband components. The beat notes between the desired carrier and the other set of adjacent channel sideband components may be included, but usually their amplitude level is low and their audio frequency so high that they are reduced to a negligible quantity by the limited receiver audio passband. The masking effect of a strong amplitude-modulated signal may serve to improve the apparent selectivity of the receiver if the strong signal is the desired one. This effect may be enhanced by the reduction of carrier beat and "monkey chatter" if the audio signal is "rolled-off" at a rather low frequency.

b. While adjacent channel interference is among the more serious problems encountered in military operations, there is little that can be accomplished in receiver design to minimize this type of interference other than maintaining the overall receiver bandwidth at the absolute minimum required to accomodate the particular type of intelligence to be received. Proper tracking and alignment of the rf stages is important. Frequently, bandwidths are made wider than necessary merely to make rf, lo and mixer tracking simple. More care in this area would greatly minimize the problem of adjacent channel interference.

3-34. Desensitization Interference

Desensitization is the resultant reduction in receiver gain for a desired signal when a strong undesired signal is simultaneously received.

a. In am communications receivers, blocking of a desired signal by interfering pulsed radar spectral components usually occurs in one of the late if stages; often the last if stage. In this stage, the interfering pulse amplitude is great and large time-constants exist in grid circuit charging networks which produce overload and pulse stretching. An indication that overload may arise is obtained from the following equation when its solution is less than unity:

$$\frac{\text{Signal}}{\text{Interference (peak)}} = \frac{R \ B_e \text{ (output)}}{1.5n \ B_e \text{ (internal)}} \qquad (3\text{-}72)$$

where:

B_e (output) = the effective bandwidth, in cps, of the receiver at the output terminals

B_e (internal) = effective bandwidth in cps, up to the point in question in the receiver

n = permissible signal to interference ratio measured on an rms basis

R = pulse repetition frequency, in pulses per second, of the pulsed radar interference

A receiver can withstand the greatest ratio of interference to desired signal when the desired signal level is such that the tubes are operating in the linear portion of their transconductance curves.

b. When intensity-modulated displays are used in radar receivers, video overload can occur in the presence of cw signals due to the limiting used with this type of display to prevent defocussing on strong signals. Use of a high-pass filter between the second detector and the video system to eliminate the dc component of the cw signal precludes the possibility of video overload.

c. Intermediate frequency amplifier saturation can also result from the presence of a strong cw signal. If the amplifier is inadequately shielded and decoupled, if oscillation may also result.

3-35. Heterodyne Interference

Heterodyne interference is caused by an off-channel signal mixing with a local oscillator frequency, its harmonics or other signals, to produce one or more heterodyne frequencies within the if pass-band of a receiver. This type of interference may be minimized by good design practices such as recommended for reduction of image and other spurious responses.

3-36. IF Rejection

Poor if rejection generally stems from failure to consider responses far out of the receiver pass-band. IF rejection figures as low as 40 or 50 db are not at all uncommon, even in receivers where other spurious responses may be well below these values.

a. In a uhf receiver, for example, it is typical for the designer to employ small values of decoupling capacitors and/or inductances which provide adequate insertion-loss in the receiver tuning range, but which, at the if frequency of say $< \frac{1}{10}$ of the tuned frequency, may provide an insertion loss of the order of 30 db less. Quite frequently, the same low values of decoupling capacitors are used in agc and power circuits entering the if strip, so that strong signals at the frequency appearing at the rf stage are not sufficiently attenuated through these circuits. Since, in the majority of vhf and uhf receivers, most of the gain is achieved in the if amplifier, a relatively small amount of signal coupled into the first if stage is sufficient to produce a response. The problem requires no extensive mathematical treatment. it is quite straightforward in that if a signal at the intermediate frequency is only attenuated 40 db between the antenna terminals and the first if grid, then the if rejection will only be 40 db down.

b. The same shielding and circuit isolation techniques, required to preclude other spurious responses, will improve if rejection. However, the major pitfall is generally that the decoupling networks used in the rf and/or if section must employ large enough values of C to provide high values of attenuation at the if, as well as the rf, frequencies. If the if amplifier is adequately shielded and ALL circuits (agc, power, etc.) entering the if module are well filtered at the if frequency then those in the rf section are of less consequence, and smaller components may be used. Feed-thru capacitors should be used in decoupling networks in if amplifiers as well as rf amplifiers if the if is in the megacycle range. Larger values of by-pass capacitors which would theoretically provide high attenuation at an if frequency of 10 mc, for example, if connected by any appreciable length of lead, may readily resonate well below that frequency and thus be totally ineffective at the frequency which it was intended to attenuate. If two-terminal (by-pass) capacitors are used, they should be of the extended foil type, with case ground and minimum lead length connecting the "high" side. (See Section II of Chapter 3 regarding suppression capacitors).

3-37. The Receiver as an Interference Source

All too often, designers fail to consider the receiver as a source of interference to other receiving devices with which it might eventually be employed in military operations. While consideration might be given to spurious responses in the design of the receiver, rarely does the designer take extensive precautions against interference signal propagation from the receiver. The seriousness of interference propagation from receivers is seldom appreciated by the designer, since it is not exhibited in his own receiver. The designer might consider spurious responses, intermodulation rejection, if rejection, etc., as serious in the receiver design, but the spurious signal propagation from the receiver, which could produce the same interference problems in an adjacent receiver as would poor spurious response rejection, is seldom considered. If one considers a countermeasure or electronic reconnaissance facility with, for example, twenty receivers, each monitoring a different frequency and each radiating local oscillator energy and its second harmonic, it can be seen that the potential interference problem to each receiver is equivalent to 38 spurious responses. The probability of frequency coincidence between other receiver lo radiations and the tuned frequency of any one receiver is just as great as the probability of coincidence of a transmitter signal and a spurious receiver response frequency. There is also the matter of detection by the enemy of spurious radiation from tactical receivers which would divulge location or mission information. The observation of "radio silence" is frequently required in tactical situations, but its purpose can be readily negated by receiver emanations. Table 3-8 contains a brief tabulation of a number of the more common sources of interference emanating from receivers, the various means of propagation, and the probable causes or deficiencies which permit, and sometimes enhance, propagation of interference from a receiver. These same deficiencies, which allow interference propagation from a receiver, contribute equally to spurious responses and/or to receiver susceptibility. Design modifications made on a particular receiver have resulted in almost the same magnitude of reduction of spurious responses and susceptibility as was achieved in the reduction of spurious signal emanations.

a. Local Oscillator Radiation. Local oscillator signals appearing across the antenna terminals of a receiver are generally the more serious, insofar as interference to other receivers is concerned. However, in areas of high concentration of receivers, such as in a forward area communications center, antennas may be sufficiently close to other receivers so that radiation from cases, wiring, etc. may cause serious interference. This is similarly true in drone and missile applications, where equipments and antennas are in close proximity with little or no shielding between antennas and wiring or equipment housings.

Table 3-8. Typical Receiver Interference Emissions

Source	Type of Interference	Propagation Media	Causes
Local oscillator and harmonics	CW	Antenna radiation	Inadequate shielding and decoupling between lo and rf stages
		External wiring radiation	Lack of filtering of power circuits into lo coupling & into adjacent wiring Poor lo shielding
		Case radiation	Inadequate shielding and filtering of lo circuits Poor overall shielding
		Conducted through common power circuits	Inadequate shielding and filtering of lo circuits and power input circuits
Intermediate frequency amplifier and harmonics	CW	External wiring radiation	Insufficient shielding and decoupling of if circuits and/or wiring entering if
		Case radiation	Insufficient shielding and decoupling of if circuits and/or wiring entering if Inadequate case shielding
		Conducted through common power circuits	Inadequate shielding of if circuitry Inadequate filtering of leads entering if strip
		Video and other output circuits	Inadequate shielding of video circuits and video wiring

Table 3-8. Typical Receiver Interference Emissions (cont'd.)

Source	Type of Interference	Propagation Media	Causes
Oscillator frequency multipliers (harmonics of fundamental and combinations)	CW	Antenna radiation	Inadequate shielding and decoupling between lo and rf stages
		External wiring radiation	Lack of filtering of power circuits into lo coupling into adjacent wiring
		Case radiation	Inadequate shielding and filtering of lo circuits Poor overall shielding
		Conducted through common power wiring	Inadequate shielding and filtering of lo circuits and power input circuits Lack of selectivity in multiplier stages
Power supplies (diode switching)	Broadband	Power input wiring radiation	Insufficient (or lack of) power input filtering
		Conducted through input power circuits	Insufficient (or lack of) power input filtering
		Power supply to receiver wiring radiation (remote power supplies)	Lack of rf filtering of power supply, output lines (or shielding of output lines)
		Case radiation	Poor shielding of power supply or power supply section of receiver
If and video amplifier (radar and pulse modulation receiver)	Broadband	External wiring	Inadequate shielding and filtering of if and/or video amplifier and entering and exiting circuits

Table 3-8. Typical Receiver Interference Emissions (cont'd.)

Source	Type of Interference	Propagation Media	Causes
		Conducted and radiated via power input	Coupling of if and video fields into other wiring
		Power supply to receiver wiring - radiation	Inadequate filtering and decoupling in if and video circuits
		Video and display circuit wiring	Lack of, or inadequate, shielding of video wiring Poor shield terminations
		Case radiation	Poor overall shielding in addition to one or more of above
Electromechanical tuning, band switching or other control functions, and remote readouts	Broadband and CW*	External wiring radiation	Insufficient shielding and/or filtering of source (contacts, motors, relays, etc.)
		Conducted and radiated via power circuits	Coupling into control wiring Inadequate shielding and/or filtering of control circuits
		Case radiation	Poor case shielding in addition to one or more of above deficiencies

*Control circuits, readout circuits, etc., entering rf or if units, also conduct rf signals out unless properly filtered over frequency range of the rf fields in the particular unit. Telemetry circuits to monitor certain circuits or functions present similar problems.

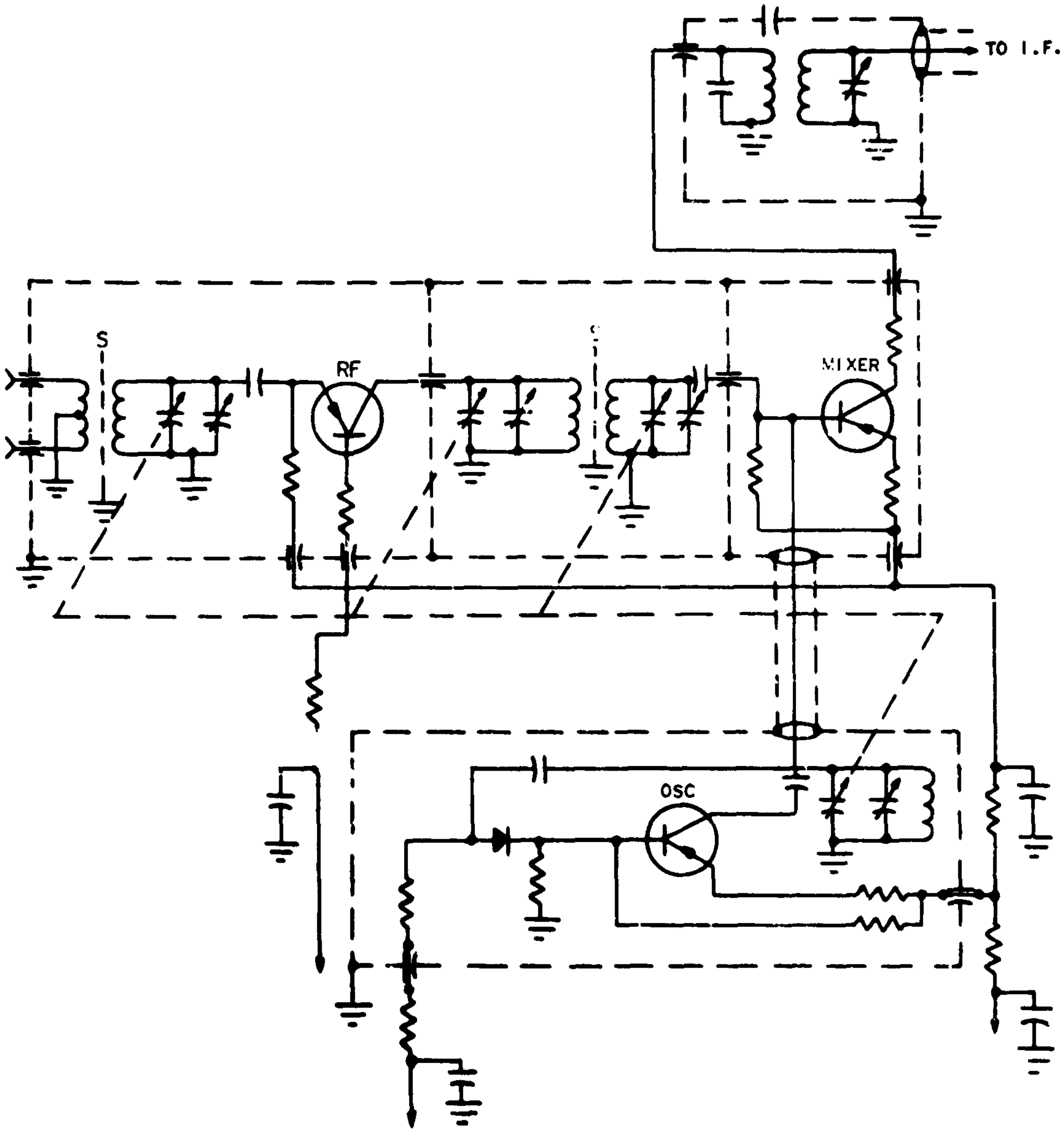
(1) Local oscillator radiation from the antenna is due to coupling of oscillator signal energy back into the first rf stage. If no rf stage is used, that is, if the input feeds directly into the mixer, the problem of isolating the oscillator signal from the antenna becomes difficult, if not impossible, since it is dependent entirely upon the Q of a single-tuned circuit. Multiple double-tuned circuits can be employed without amplification, but would introduce losses.

(2) In variable-tuned receivers, it becomes quite difficult to provide sufficient attenuation of the local oscillator signal without use of rf stages. However, in fixed-tuned receivers it is possible to employ sharp band-pass filters ahead of the mixer, a lo rejection filter, or a low-pass or high-pass filter, depending upon which side of the tuned frequency the oscillator operates. In any case, careful shielding must be employed, since no filter is any more effective than the shielding provided between its input and output. The same principle applies to tuned circuits in each rf stage, and the oscillator circuitry as well. The fields around the oscillator coils, and all associated circuitry including those around the tube or transistor, must be confined by adequate shielding. Each power and/or control lead entering the local oscillator enclosure must be decoupled, using either R-C or L-C networks. Where dc voltage drop is a factor (and where space permits), L-C filters should be employed. In the uhf range and above, inductances may become resonant because of distributed capacitance, and RC networks may be necessary. Resistors must be chosen on the basis of their rf characteristics also. Film types, or thin "stick" resistors with minimum shunt capacitance, must be employed.

(3) The same techniques of confinement and filtering employed throughout the lo, mixer and rf stages will reduce lo radiation

from the case and external wiring. These measures, combined with low local oscillator power and good oscillator waveform to reduce harmonic content will, when properly employed, substantially reduce the problem of local oscillator emissions from the receiver. These same techniques, combined with careful design of the tuned circuits, are also necessary and will also be quite effective in reducing spurious responses. Figure 3-143 shows the rf section of the same receiver depicted earlier in figure 3-126 with design modifications included to minimize local oscillator radiation and spurious responses.

b. Intermediate Frequency Signals. Radiation and conduction of if signals from a receiver is probably one of the most frequently ignored areas and one which has been found to produce serious compatibility problems in several complex receiving systems. In one case, a 60-megacycle if signal was found to exist on almost completely unrelated circuits within the system, and was causing interference problems in certain automatic functions. Levels, equivalent to 1000 μv/m and 100 μv/m at 60 and 120 megacycles (2nd harmonic of the same if) respectively, were observed to be radiating from several points within the system. This was attributed to poor shielding of housings and wiring, and inadequate filtering of power and other circuits entering the if amplifiers. The decoupling employed to prevent feed-back or oscillation does not necessarily prevent if signals from being conducted out of the latter stages and coupled into other wiring, or simply radiating from a poorly shielded housing. Neither does it necessarily preclude entry of signals at the if frequency into the early if stages, causing susceptibility at the if frequency. The decoupling CAN, however, be so designed as to serve both functions with very little additional effort and few additional components. If interstage decoupling is to be employed internally within the if chassis, (in a B+ circuit for example), then the power feed should not enter at an early (high gain) or last (high signal level) stage, but rather at an interme-



S = ELECTROSTATIC SHIELDS

= FEED-THRU CAPACITORS MOUNTED THROUGH CHASSIS OR SHIELD

ALL GROUND LEADS AND BY-PASS CAPACITOR LEADS OF MINIMUM LENGTH

IN1212-331

Figure 3-143. Typical VHF Receiver Design to Minimize LO Interference and Spurious Responses diate stage, at which point neither sensitivity nor signal level are maximum (see figure 3-144). This simplifies the decoupling network and takes advantage of the interstage decoupling required, in any case, for proper operation. If interstage decoupling is not employed within the if chassis or module, then the decoupling methods shown in figure 3-145 should be employed.

(1) L-C decoupling, as illustrated in figures 3-144 and 3-145, is necessary in filament circuits; while in agc circuits, R-C networks are more appropriate since voltage drop is not a factor. In any case, each lead entering the if chassis or enclosure must be decoupled by either L-C or R-C networks. This includes if gain control if used to control if bias.

(2) The output of the second detector must also be carefully filtered to attenuate the if signal energy, whether the output is a video or audio signal. The high-level if output signal can be coupled around or through the detector and be conducted out to external wiring, as well as coupled into power, audio, or other circuits which might not be shielded as well as the if amplifier. Therefore, careful filtering at this point is important.

<u>c.</u> <u>Power Supplies and Electromechanical Devices.</u> Interference reduction techniques for power supplies, relays, and other electromechanical devices are covered in considerable detail in Sections III and IV of this chapter. These methods apply equally to receiver power supplies, control circuitry, relays, etc.

(1) If the source suppression measures heretofore discussed with regard to containment of receiver signals are employed, then additional precautions to preclude rf signal propagation through the power supply is unnecessary. Filtering can then be designed to cover only the frequency spectrum of the power supply diode

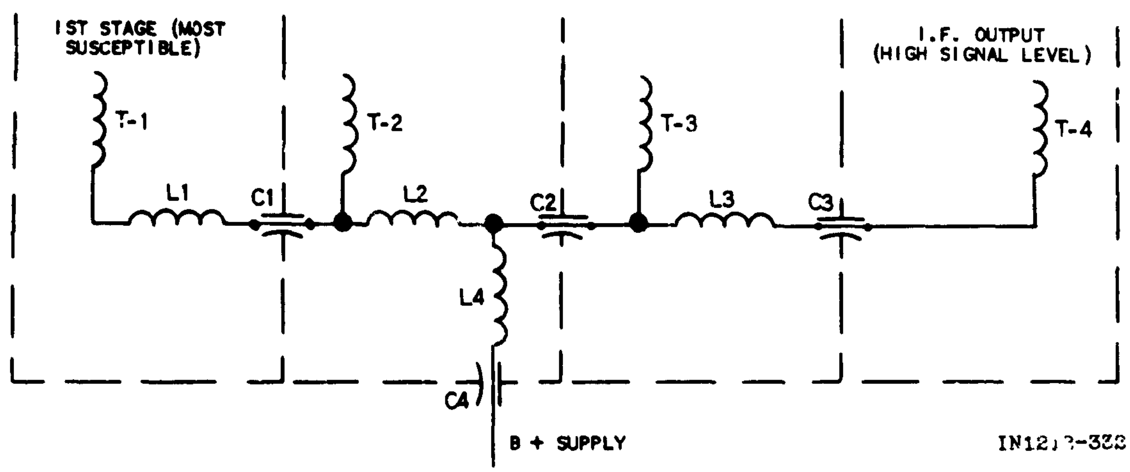


Figure 3-144. Typical IF Decoupling

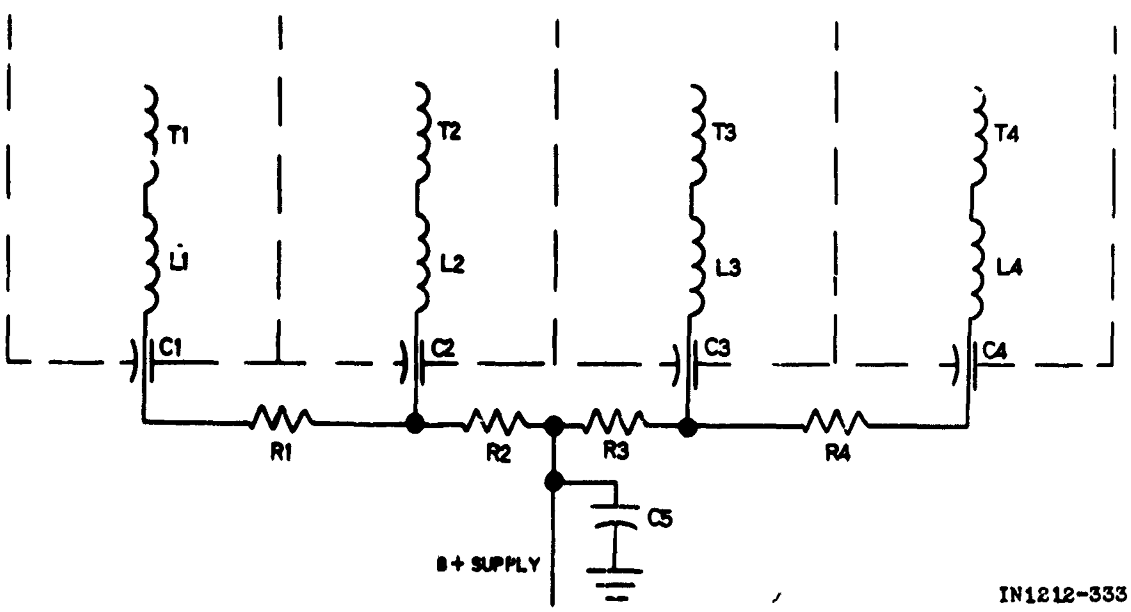


Figure 3-145. Typical Alternate IF Decoupling

switching pulses.

(2) If these design features are not employed, then it must be assumed that any signals present within the receiver will be coupled back into the power supply and into the input power circuits to be radiated or conducted into other equipment. It is commonly believed that solid-state power supplies provide rf isolation; however, this is rarely the case. Therefore, if rf signals (lo, if, etc.) are allowed to propagate within the chassis, then power supply input filters must be used which are effective up to the highest frequency of the signals present within the receiver housing. Under these conditions, the overall receiver housing must be a complete rf-tight enclosure to prevent radiation.

(3) One particular electromechanical device often employed in receivers is considered worthy of specific mention -- the thermostatically-controlled oscillator crystal oven. These devices are prolific sources of broadband interference and also a source of coupling of the oscillator signal out of the oscillator module or enclosure. Not only is this a source of interference to other equipments, but has often been found to produce interference within the receiver in which it is contained. Not only must the switching transient be suppressed at the thermal switch itself, but the heater conductors must also be filtered adequately to suppress the oscillator signal which would otherwise be coupled out on the heater lines.

(4) Proportional type heaters (rather than the thermal switch type) are available which eliminate the switching transient and are highly recommended for this purpose. However, it must be remembered that, although not itself a source of interference, the heater circuit is capable of conducting oscillator signal energy out into chassis wiring.

3-38. PRF Discrimination

When a victim radar is exposed to pulse interference from another radar or pulse source (a "culprit") whose pulse repetition frequency is nearly equal to a multiple or sub-multiple of its prf, a "running-rabbit" type of interference pattern appears on the victim's ppi. Typical patterns are shown in figure 3-146A, where the victim's prf is just below the culprit's ($f_v < f_c$), and in figure 3-146B, where the victim's prf is slightly higher than the culprit's ($f_v > f_c$). Sometimes, radars are operated in close proximity so that complete elimination of the interference is not feasible. By closely examining "running-rabbit" patterns and applying nomographs that describe running-rabbit parameters, a victim can determine the prf of the culprit, thus obtaining a clue to the culprit's identity.

a. Pulse Delay. One method of eliminating pulses with repetition rates different from a victim radar is to delay each incoming pulse by the repetition period of the radar and allow it to be displayed only if it coincides with the next incoming pulse. A simplified block diagram for accomplishing this is shown in figure 3-147.

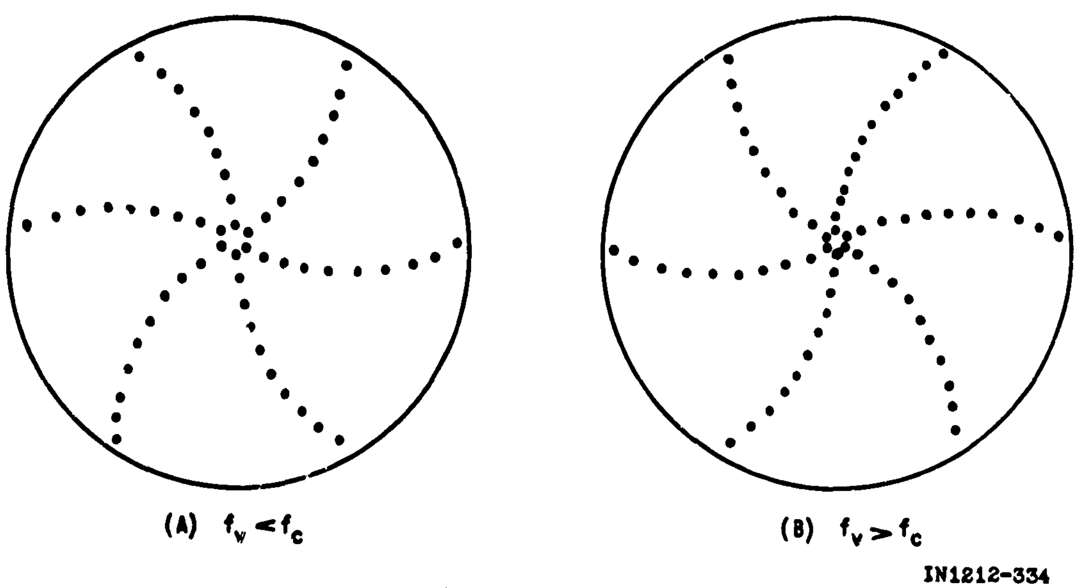


Figure 3-146. Running-Rabbit Patterns

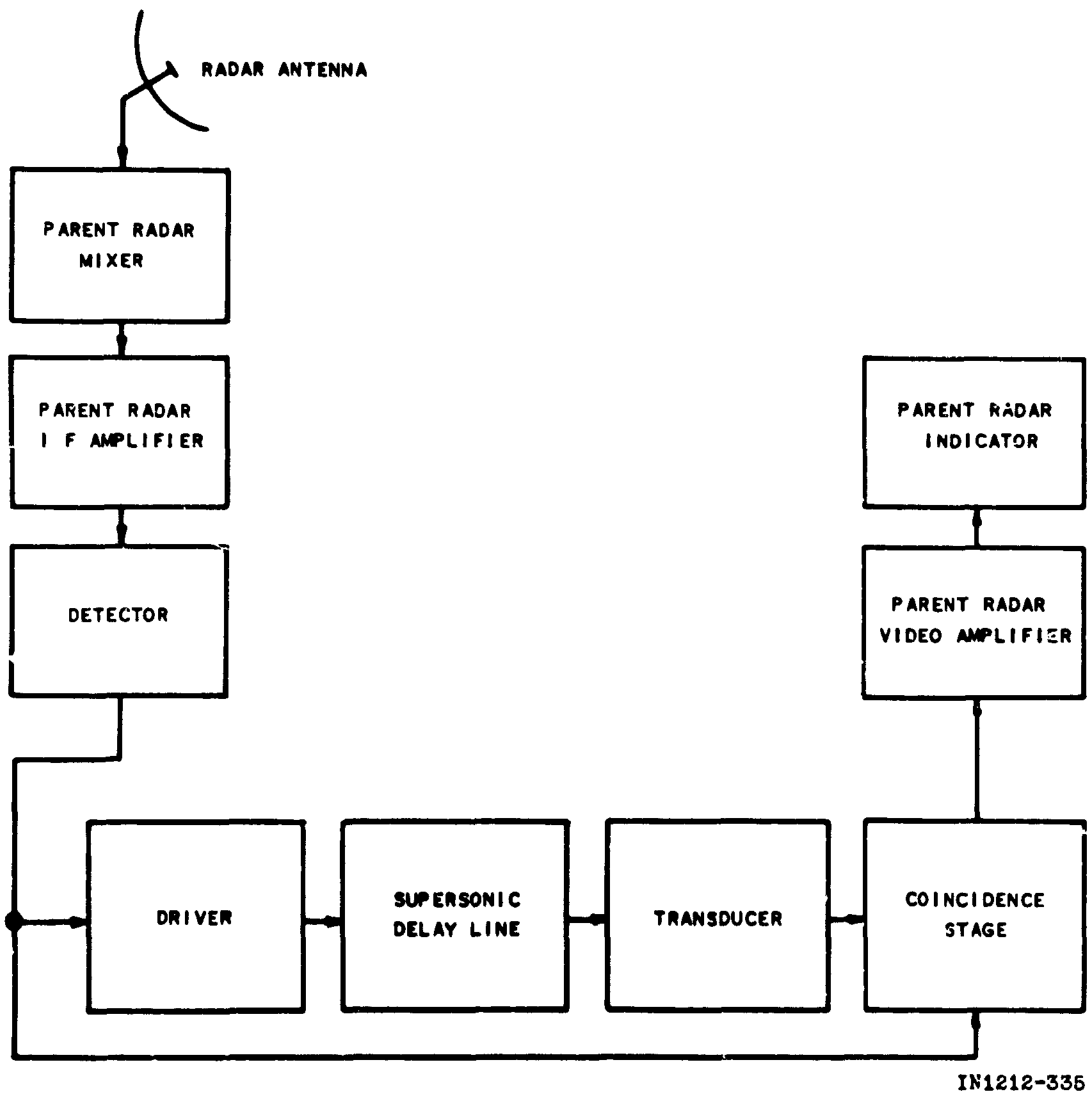


Figure 3-147. Simplified Block Diagram of a PRF Discriminator Circuit

(1) A detected incoming pulse envelop modulates a 10-mc oscillator which drives a supersonic delay line. The material and length of the line determine the time delay of the pulse as it travels down the line. For example, with a repetition rate of 1050 pps, the delay would be equal to $\frac{1}{1050}$ = 952 microseconds. The delayed pulse and the succeeding undelayed pulse then enter the coincidence stage. Only if both pulses arrive at the same time are they allowed to pass to the video amplifier and indicator. Pulses, whose repetition periods do not equal 952 microseconds, will not arrive together at the coincidence stage and, therefore, will not enter the video amplifier.

(2) In cases where the parent radar has a staggered prf, a delay line for each prf can be used, and the proper delay line switched into the circuit in synchronism with the radar's prf generator.

(3) Other useful prf discriminator circuits exist. Some of these are: video integration and slowed-down video. A simple block diagram of a video integration circuit is shown in figure 3-148. The integration method utilizes supersonic delay lines to store target returns, which are then fed back to the input of the delay line where they add to succeeding incoming pulses. This continues until a desired threshold is reached, whereupon the pulses pass through the amplitude selector to the video amplifier. Pulses of incorrect repetition rate will not coincide and add, and therefore, will not build up to the threshold level. Slowed-down video also uses a delayed video process, but matches a number of successive pulses for time coincidence. A definite, fixed quantity of coincidences (up to 8) can be required before the signal is allowed to pass to the indicator.

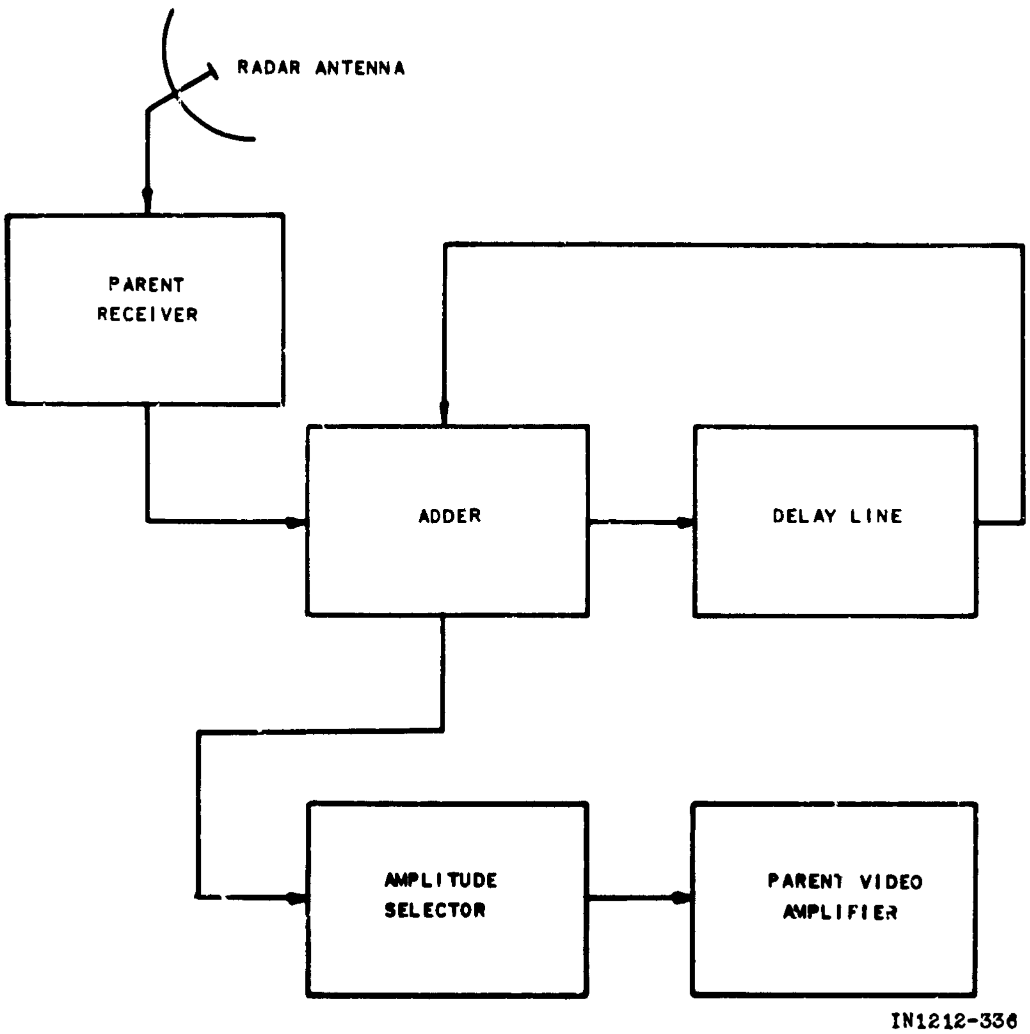


Figure 3-148. Simplified Video Integration Block Diagram

b. Double Threshold. The "double-threshold" method of detection uses prf discrimination. A certain fraction of expected target echoes is required to exceed the receiver threshold at the "precise" pulse repetition frequency before a target is declared present. For example, 16 echoes might be expected from a target as the main beam sweeps past, and the second threshold might be set at 8 pulses received at the precise transmitter prf. "Precise" in the above statements can be taken to mean that the prf of the accepted signals cannot deviate from that of the transmitter by more than 0.01 or 0.02 per-cent. Otherwise, the incoming pulses would "walk through" the acceptance time intervals before the criterion of filling the 8 gates was satisfied. In video or post-detection integration, supersonic delay lines are used to store target echo pulses. These pulses are fed back to the video input and added until a desired threshold is attained. If the time of arrival of successive pulses does not correspond to the interpulse period, no integration takes place and a target is not declared to be present.

c. Random PRF Emission. Deliberate prf jitter, in conjunction with receiver gating, can aid in avoiding other emissions with relatively stable prf's. In other words, if periods between pulses are staggered in a random or near-random fashion, and receiver gates are set to be open during these intervals, the probability of a stable prf entering the gates often enough to cause degradation is extremely small. The interference condition that such prf identification techniques are not designed to overcome is that of a nearly identical interfering prf. For this case, it is most desirable to have a choice of operating prf's available to the operator. Even then, it may not be possible to find a clear prf in a dense signal environment. Also, the equipment for prf discrimination techniques is usually bulky and expensive. The techniques usually result in some range reduction. Therefore, prf discrimination is not a cure-all and should be used in conjunction with other methods in high-density environments.

d. <u>Pulse-Width Discrimination.</u> Two types of pulse-width discriminators are generally used. One type, known as an integrator-discriminator, generates a ramp which continues to rise for the duration of a delayed incoming pulse. If the pulse is wide enough, the ramp reaches a preset threshold level and a gate allows the original pulse to pass on to the indicator circuits. In this way, the circuit discriminates against pulses of less than a specified duration. Figure 3-149 shows a block diagram of a circuit which accepts pulses whose durations lie within definite lower and upper limits.

(1) The detected video pulse from the parent radar controls the duration of a ramp produced by the ramp generator. If the pulse is wide enough, the ramp passes through the amplitude selector to produce a trigger. This trigger actuates the gate generator, which turns on the gated video amplifier. The receiver pulse, which was delayed awaiting the above pulses, then passes through the video amplifier to the indicator. If the detected video pulse is wider than the parent radar's pulse, the ramp continues to fall at the output of the cathode follower. It reaches the preset threshold of the second amplitude selector and then nullifies the effect of the trigger on the gate generator; thus, no gate is produced. The gated video amplifier remains shut-off, and the wide pulse does not reach the indicator. The setting of the amplitude selectors in this circuit determines the minimum and maximum pulse widths that will be accepted.

(2) Another type of pulse-width discriminator uses delay lines, generally as reference standards, to which the width of the incoming pulse is compared. A simplified block diagram of a circuit which rejects pulses with durations outside fixed low and high limits is shown in figure 3-150. Its operation is explained by using the waveforms shown in figure 3-151. A desired pulse

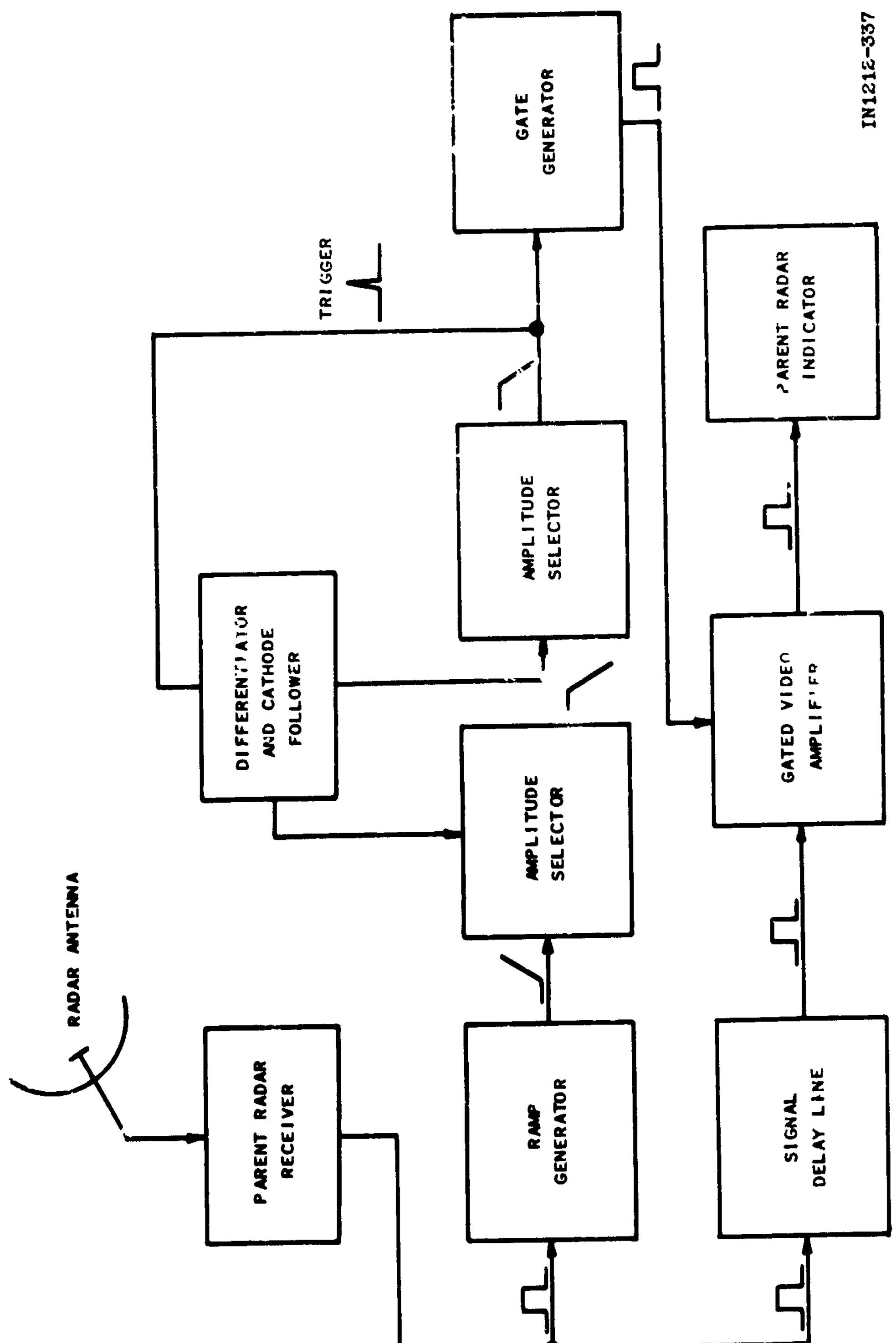


Figure 3-149. Block Diagram of Integrator Type Pulse-Width Discriminator

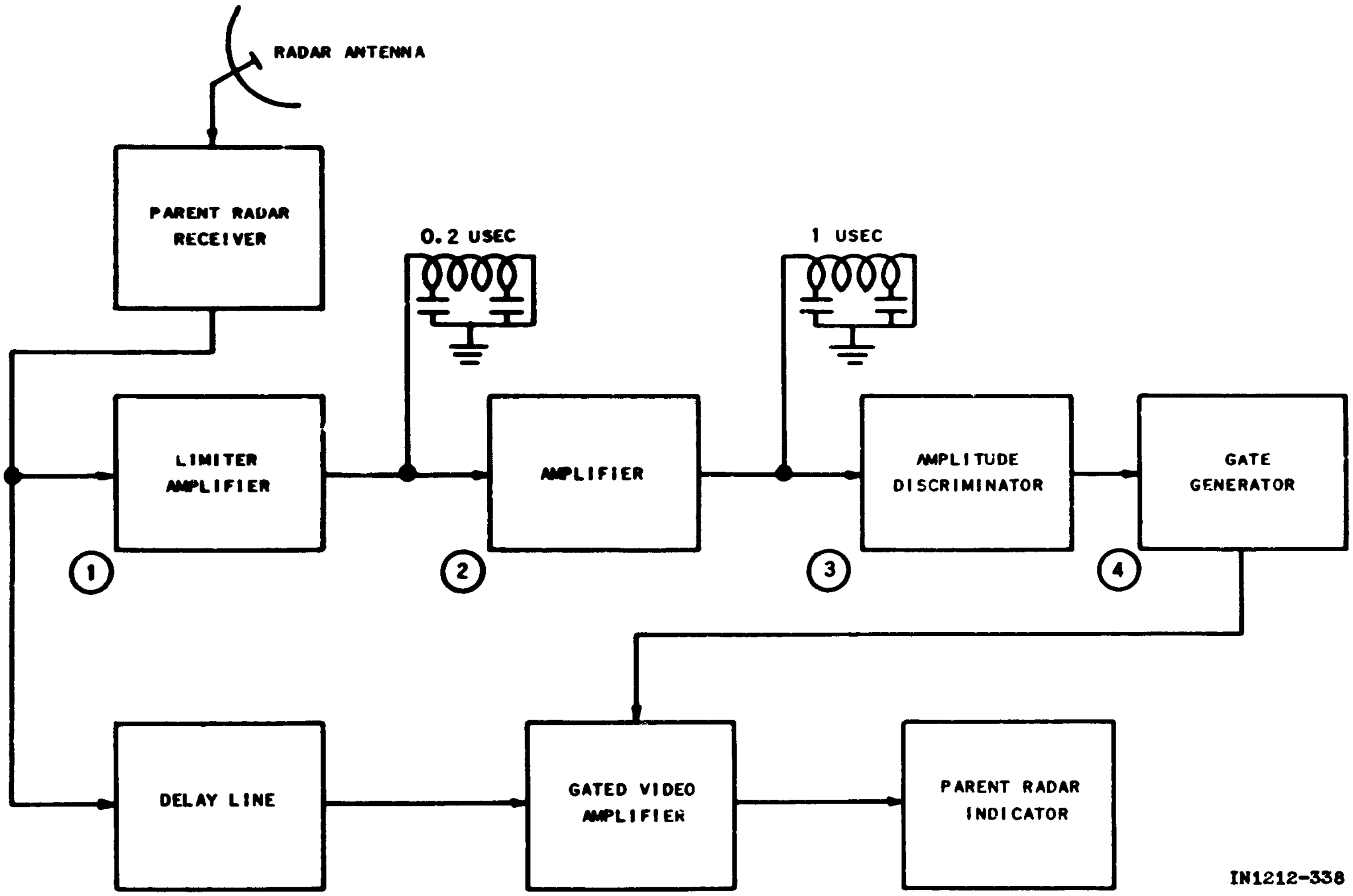


Figure 3-150. Delay Line Pulse-Width Discriminator

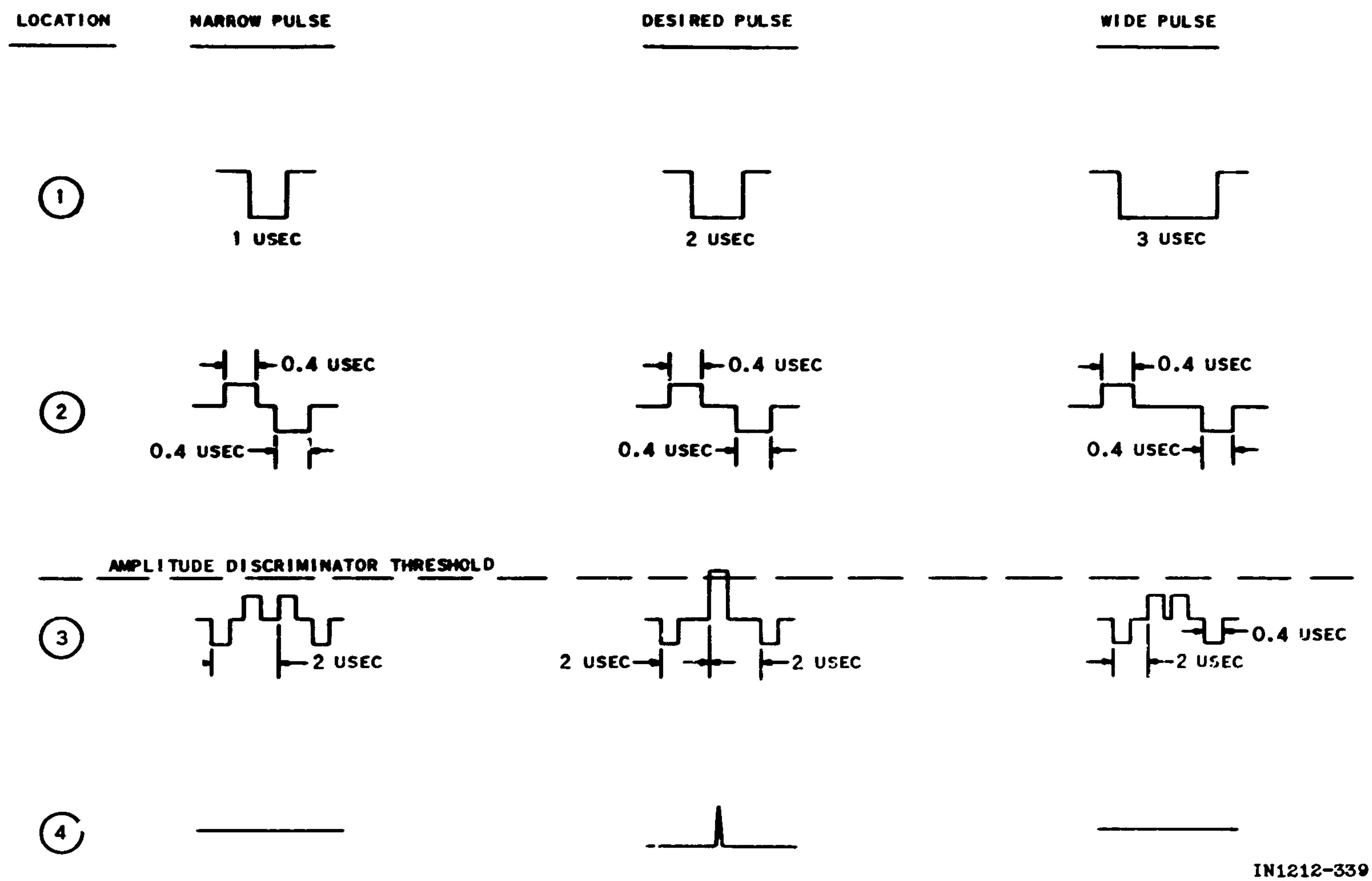


Figure 3-151. Waveforms for Pulse-Width Discriminator

width of 2 microseconds will be assumed. The 0.2-microsecond delay line, shorted at its output, transforms an input pulse to two pulses, each of 0.4 microseconds, twice the duration of the delay line. The leading edge of the first pulse starts at the leading edge of the receiver input pulse. The second pulse starts at the trailing edge of the receiver input pulse. These pulses are inverted by the amplifier and transmitted down the 1-microsecond delay line. They reappear at point (3), inverted, 2 microseconds later. Of the three pulse-widths shown, only the 2-microsecond input pulse overlaps at point (3) to build up to the amplitude discriminator threshold. This, in turn, triggers the gate generator, which turns on the gated video amplifier, allowing the 2-microsecond pulse to pass to the indicator.

3-39. Instantaneous Automatic Gain Control

A situation that arises often enough in radar systems (and possibly also in other pulse applications) to demand special consideration is that in which interference in the form of a cw carrier or an amplitude-modulated wave is present. If the amplitude of the interference is small compared with the desired pulse signal, it is ordinarily sufficient to provide adequate gain control. Where the relative amplitudes are reversed, however, additional design features may be useful. These features act in such a way as to preserve the incremental gain of the receiver for the desired pulse, while greatly reducing the response to the interfering signal.

a. Suppose that the desired signal is made up of a carrier frequency, amplitude-modulated by a rectangular pulse. It is convenient to focus attention on the if amplifier, and therefore f_o may be assumed to be in the neighborhood of an intermediate frequency, say 30 mc/sec. For ordinary pulse lengths, such as 1 μsec, the signal then consists of approximately 30 oscillations. Suppose, also, that the interfering signal is a cw carrier of frequency f_1. If $(f_1 - f_0) > 1$ mc/sec, the envelope of the combination -- signal plus interference -- will exhibit at least one complete cycle of the difference frequency $(f_1 - f_0)$. If, however, $(f_1 - f_0) << 1$ mc/sec, the envelope of the combination will be essentially the steady value characteristic of the cw carrier with a rectangular pulse superimposed on it. According to the relative phase of signal and interference, the resultant may be either larger or smaller than the cw interference; therefore, the envelope may appear to have either a positive or a negative pulse of amplitude equal to that of the signal, or there may be any intermediate condition. If repeated pulses are viewed on an oscilloscope in the presence of such interference, all values of the relative phase between signal and interference are run through, and the pattern appears to be filled in as shown in figure 3-152B. It is apparent, therefore, that for part of the time at least, there is a change in the amplitude of the envelope of signal plus interference, which is comparable in magnitude to the amplitude of the signal alone. If this change in envelope can be preserved, the signal may be recovered from the interference.

b. The phenomenon of if overloading is well known. It will be recalled that if the input vs. output characteristic of a single if stage is measured, a level of input signal is easily found such that increase in input produces no increase in output. If this stage is so placed in the receiver that the level of the interfering carrier at its grid is in this region of no change of output, it is plain that the signal pulse superimposed on the carrier envelope will be entirely wiped off and the signal will be lost.

c. Two procedures might be adopted to combat this signal loss. The first, reduction of gain ahead of the point at which overloading occurs, is effective, but has the disadvantage of reducing the size of the desired signal at the receiver output. The second procedure is provision of back-

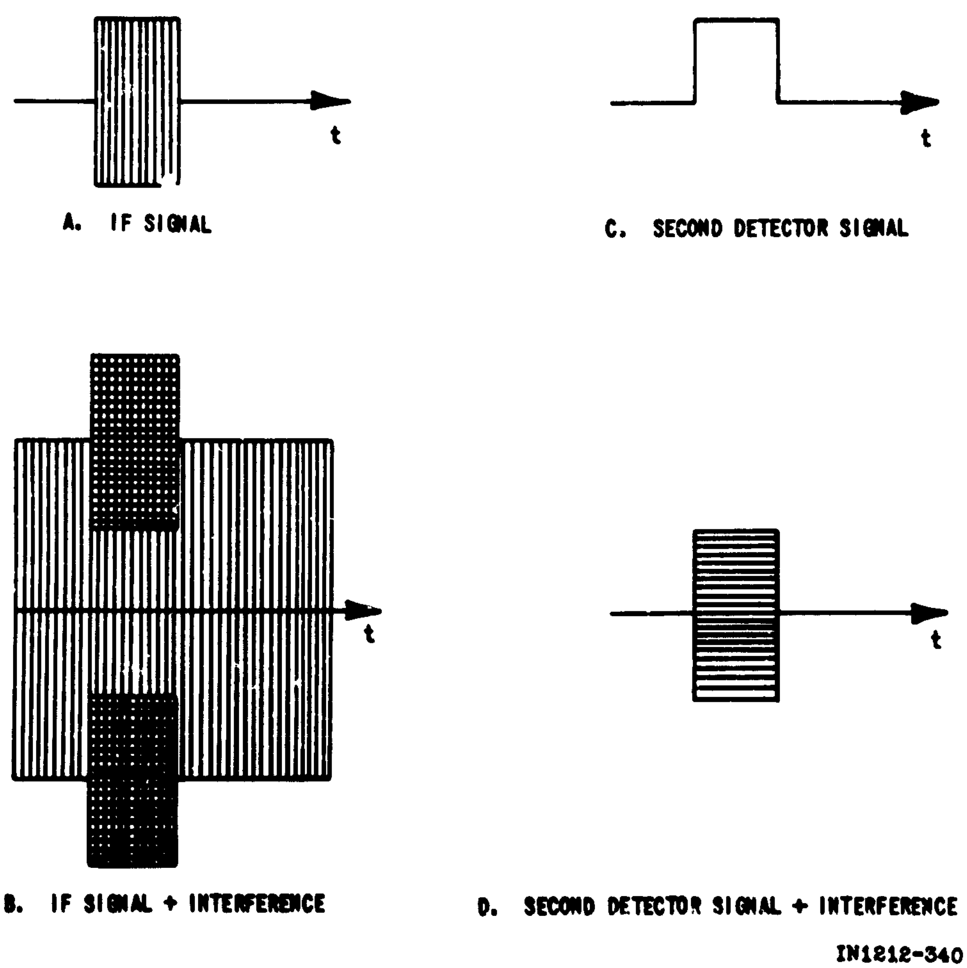


Figure 3-152. Waveforms of Signal Plus Interference

bias in some form. An additional bias voltage is added between grid and cathode. Ideally, the magnitude of this bias should be comparable to the peak voltage, or envelope, of the interfering carrier at the grid of the tube in question. In this case, the peaks of the waves of the interference come to approximately the normal quiescent point of the stage, and changes in the size of the peaks are preserved.

d. Probably the simplest circuit that may be considered for this procedure is obtained by increasing the size of the normal cathode resistor to the point where the tube acts as an infinite impedance detector. To prevent excessive loss of gain, the resistor is bypassed for the intermediate frequency. Normal operating current in the tube may be obtained by returning the cathode resistor to a negative voltage, or the grid circuit to a positive voltage. This simple circuit suffers from two disadvantages. First, it is difficult to secure enough bias voltage. In order that the bias be equal to the peak voltage of the interfering carrier, it would be necessary to achieve unity detection efficiency. A large value of cathode resistor and bypass condenser would then be required. To obtain normal gain from the stage in the absence of interference, normal values of cathode current must be used, thereby setting a limit to the size of the cathode resistor. The second difficulty arises from the undesirable time-constant of the cathode resistor and bypass condenser. If a strong signal, such as a block of normal echoes, ends suddenly, the cathode bypass condenser will be left charged to a positive voltage which it can discharge only through the cathode resistor, since the tube will be cut off. This discharge may take several microseconds, during which time the receiver is insensitive to weak signals.

3-40. Limiters

Limiters, as their name implies, are primarily restrictive devices and it can be expected that distortion will result from their use, particularly when the input exceeds the limiting threshold. Limiters of the instantaneous noise-peak type generally distort the output whenever the modulation of the incoming signal exceeds a definite percentage. The effects of this distortion can be intensified by the transient distortion characteristics of the audio amplifier. In general, it is very desirable that either triode tubes be used in the audio amplifier, or that a degree of degenerative feed-back be provided which will be great enough to prevent "hang-over" or oscillations due to insufficient damping of the output circuit.

a. A good limiter of the instantaneous noise-peak type should cause little or no distortion below the modulation percentage at which its limiting action begins. Above this point, it should cause a very considerable increase in distortion with increase in modulation depth due to effective limiting of the output wave on one-half cycle. Distortion, caused by output limiters in either rf or af amplifiers, may be considerably reduced by the filtering action of tuned circuits, or by low-pass filters following the limiter. These remedial measures are generally useful, however, only for cw reception.

b. Peak-noise limiters may be of either the shunt or series type. In general, it has been found that the series type of limiter is more readily applicable to existing equipments. The series limiter is one of the simplest types of noise reduction networks. It is also one of the most useful. It is usually used in the af section of the receiver between the second detector and the first af amplifier. The basic series limiter and two modifications are presented.

(1) Basic series limiter. The basic series limiter circuit is shown in figure 3-153. The limiter operates as follows: A dc rectified voltage appears across R_1 and R_2 as a result of the if carrier. The cathode of D_2 is then negative with respect to ground by the amount of this voltage. The plate is positive with respect to the cathode by one-half this voltage.

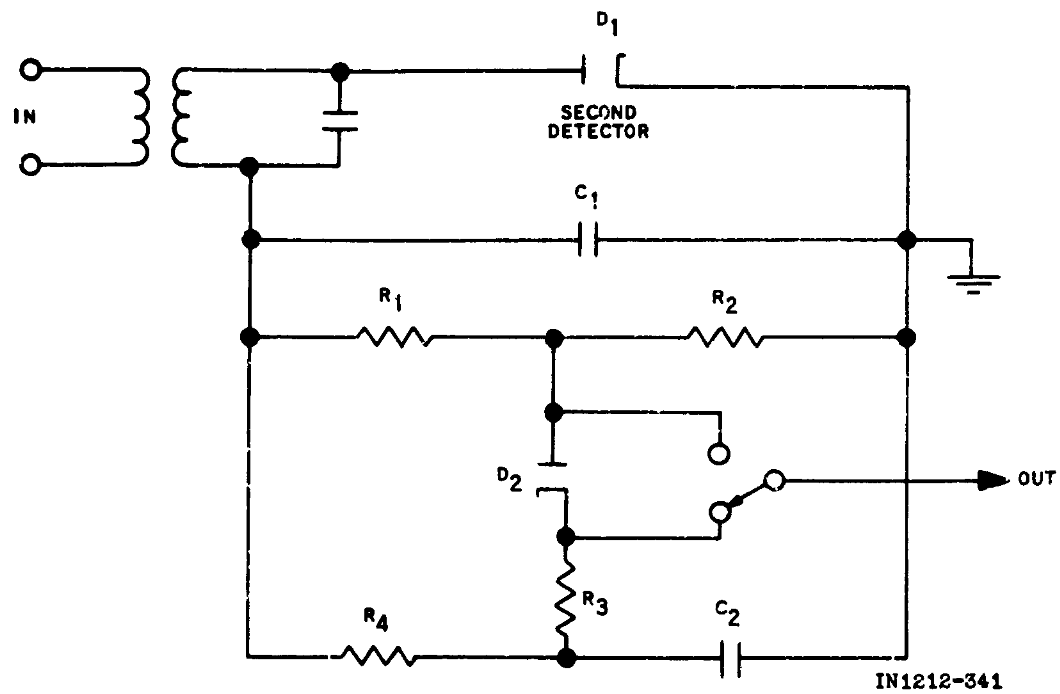


Figure 3-153. Basic Series Limiter

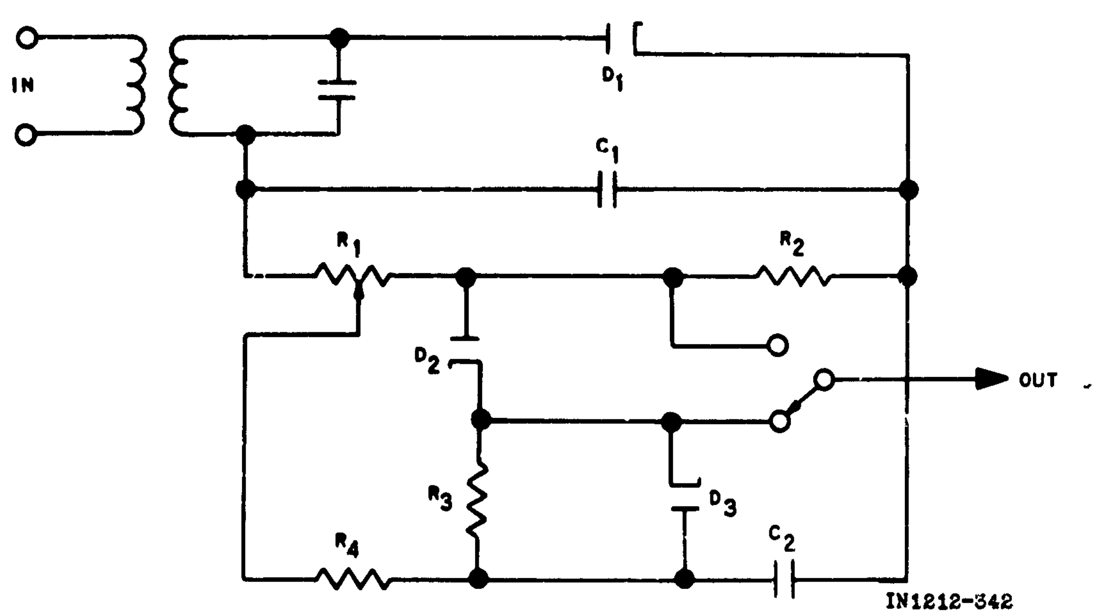


Figure 3-154. Modification of Basic Series Limiter

This causes the diode to conduct since its resistance is low compared to the rest of the circuit. The output coupling capacitor is then connected to the junction of R_1 and R_2 through conducting D_2. This allows the af signal to be fed from the detector D_1 to the af amplifier. Under these conditions, C_2 is charged through R_4 to a point slightly more negative than the plate. The time-constant of this circuit is long compared to the time-constant of the plate circuit; therefore, if a high-level noise pulse is received, it will drive the plate negative before the cathode can follow, and the diode will stop conducting. This then limits the pulse. The pulse is usually not of long enough duration to allow the cathode to go to its full negative potential; the plate will then assume a positive voltage with respect to the cathode and D_2 will conduct again.

(2) Modification of basic series limiter. Figure 3-154 is a modification of the basic circuit in that it has a potentiometer substituted for fixed resistor R_1. This allows the limiter threshold to be varied from about 65% modulation down to zero. This is done by changing the voltage of the cathode with respect to the plate of D_2. This circuit also contains another diode, D_3, shunted across R_3 to buck the internal potential of the limiter diode. This potential reduces the effectiveness of the limiter at low carrier levels by raising the limiting threshold. The voltage across R_3 will keep the diode at cutoff except at weak signals. This diode does, however, add some distortion and may be removed if distortion cannot be tolerated.

(3) Low-loss modification of basic series limiter. A low-loss modification of figure 3-153 is shown in figure 3-155. The plate and cathode connections of D_2 are reversed. The cathode is biased negative with respect to the plate by a connection to the junction of R_2 and R_3 through R_4 and R_5. R_4, C_2 is the long

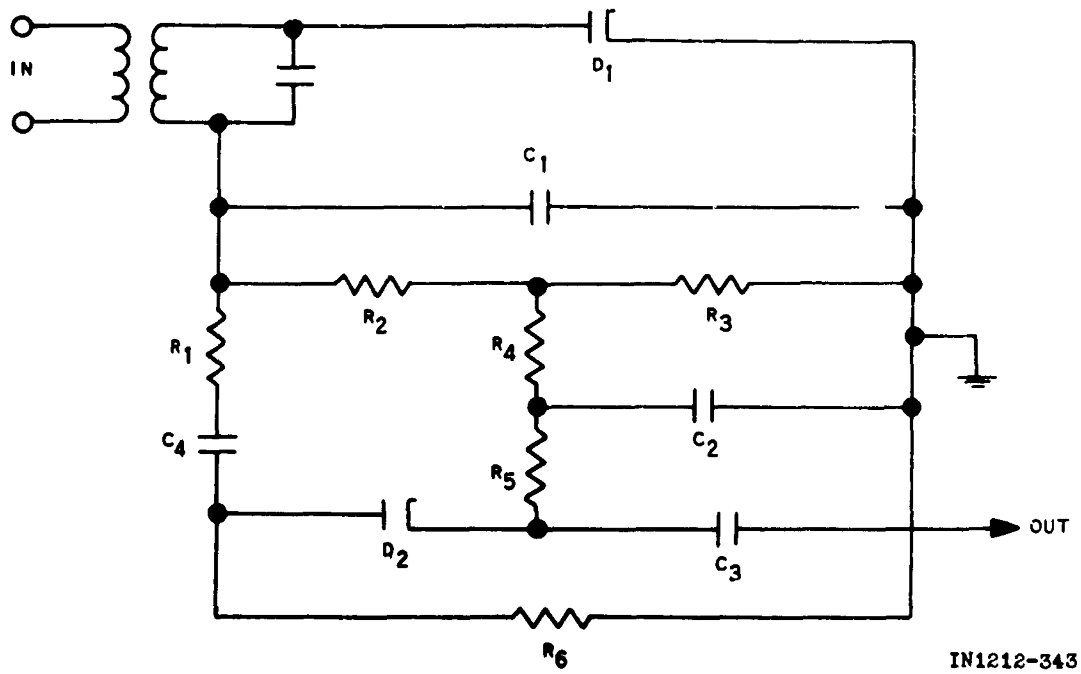


Figure 3-155. Low-Loss Modification of Basic Series Limiter

time-constant cachode circuit as in figure 3-153; and R_5 is the cathode load resistor for D_2. R_1 is an rf filter element. The limiting action of this circuit is identical with that of figure 3-153.

(4) Summary. The limiter with its modifications is effective in reducing the high-level pulse interference. Figure 3-153 has the disadvantage that it only allows about one-half of the output of the detector to be fed to the af amplifier. This affects only the reserve gain, and not the sensitivity of the receivers. As previously mentioned, diode D_3 of figure 3-154 will allow limiting at lower carrier levels, but does add some distortion to the audio output. Figure 3-155 allows more of the signal to be fed to the af amplifier than figure 3-153, but like figure 3-153 has no means of controlling the limiting threshold.

c. Interfering pulses frequently can be of the same order of intensity as desired pulses, and limiting circuits can reduce (or partially discriminate against) large undesired pulses. When the relative magnitude of the anticipated signal is known, techniques exist to discriminate against much stronger and much weaker signals. Sensitivity, time control, and automatic overload control are other examples of this approach.

Section VI. TRANSMITTERS

3-41. General

Transmitters, in general, are not susceptible to interference; they may, however, produce considerable spurious interference. The interference generated in a transmitter is usually categorized as spurious and harmonic emissions from the case, leads and antennas; transmitter noise; sideband splatter; and intermodulation. These and other categories of transmitter interference are described, as are optimum transmitter design and interference suppression data.

3-42. Transmitter Interference

Interference arises when off-channel transmitter signals produce on-channel responses at a receiver tuned to a frequency other than the transmitter fundamental emission frequency (which is defined as the center frequency of an assigned channel). Fundamental emissions are those rf emissions within the assigned channel; spurious transmitter emissions are any rf emissions appearing at the transmitter antenna at frequencies outside the assigned channel; and harmonic emissions are spurious rf emissions that occur at an integral multiple of the fundamental emission frequency. Spurious emissions from transmitters can be categorized as self-spurious (emissions arising from internal sources within the transmitter), and external spurious (emissions arising from the presence of interfering signals from other transmitters). There are five primary sources of spurious signals in transmitters. They are: low-level oscillator and multiplier stages with frequencies that are fractional multiples of the carrier frequency; power amplifiers with frequencies that are harmonics of the carrier frequency; mixer, or frequency synthesizing stages, with frequencies that are not directly related to the carrier frequency; parasitic oscillations in any stage; and noise. Spurious voltages, generated in a particular stage, such as a multiplier or power amplifier, are primarily dependent upon:

1) The nonlinear operation of components where current fluctuates: for example, electron tubes, transistors, and diodes

2) The impedance presented to the plate of a tube by passive circuitry

3) The input voltages to the grid of a tube, including all spurious voltages from the previous stage

a. Direct Emission. Direct emission is the emission of electromagnetic energy at any frequency and from any location on the transmitter other than the antenna; this includes transmitter-housing (case) emission as well as feeder-line emission. Radiated interference can emanate from any discontinuity in the transmitter housing; conducted interference can propagate along any conductor passing into the transmitter.

b. Harmonic Emission. Harmonic emission is emission at a radio frequency that is a fractional or integral multiple of the frequency of the fundamental, or base-frequency, oscillator. Undesired harmonic and subharmonic frequencies in the transmitter output spectrum are generated in some nonlinear element in the transmitter itself (generally in the output stage). Any distortion of the ideal sinusoidal wave-shape produces harmonics that are undesirable from the interference standpoint and represent definite power loss and lowered efficiency. The fundamental frequency, and fractional or integral multiples of the fundamental, may appear in the output spectrum. The most common off-frequency signals that cause interference are harmonics of the basic oscillator frequency. In cases where magnetrons are the basic oscillators, there may be several spurious frequencies that are not related to the carrier frequency. A common practice in transmitter design is to employ an oscillator of relatively low frequency and then multiply this frequency using one or more multiplying stages. During this process, harmonics of the base frequency are also multiplied. The frequencies of these spurious signals (f_s) can be determined as follows:

$$nf_o = Nf_o$$

where: n = the order of the harmonic of the oscillator frequency, f_o

N = the number by which the base frequency has been multiplied

c. Parasitic Emissions. Parasitic emissions are emissions at radio frequencies that are not harmonically related to either the fundamental frequency or the intermodulation products. Parasitic emissions occur when a circuit is excited and goes into oscillation at a frequency other than the desired one. These emissions include shock excitation due to internal transient phenomena and excitation of circuit components due to the carrier signal.

d. Splatter.

(1) Splatter. Splatter is emission, falling outside the assigned channel, that results from the modulation of a carrier wave.

(a) Angular modulation splatter. Frequency and phase modulation systems theoretically generate an infinite number of side-bands. The significant side-bands (those with amplitudes greater than -40 db, referred to the unmodulated carrier) increase with the modulation index. The modulation index of a frequency-modulated signal is directly proportional to the modulating signal amplitude and inversely proportional to its frequency. The number of significant side-bands decreases as the modulation signal frequency increases, assuming a constant amplitude signal input. The bandwidth, occupied by the significant side-bands of a frequency-modulated signal, is equal to twice the frequency of the highest side-band. This frequency is given by the number of significant side-bands, multiplied by the modulating signal frequency. Since the number of significant side-bands decreases with increasing frequency, the bandwidth of the frequency-modulated signal tends to remain constant. The number of significant side-bands in a phase modulation system is independent of the modulating signal frequency. The bandwidth, occupied by the side-bands, increases linearly with increasing modulation frequency for a constant amplitude signal input. In either case, the designer must select parameters that will ensure containment of all significant side bands within the assigned channel.

(b) Amplitude modulation splatter. When the amplitude of an rf output is not directly proportional to the amplitude of the modulating wave, the modulation envelope is distorted. This distortion causes new frequencies to be generated, that combine with the carrier to form new side-bands. Side-bands, caused either by distortion of the modulated wave or by nonlinear operation, widen the spectrum occupied by the modulated signal. When an rf output does not have an envelope directly proportional to the amplitude of a modulating wave, side-band distortion and splatter are produced. The output waveform, $M(t)$, for sinusoidal plate modulation is of the form:

$$M(t) = \left[1 + b_1 m \cos \omega_m t + b_2 m^2 \cos^2 \omega_m t + b_3 m^3 \cos^3 \omega_m t \ldots + b_n m^n \cos^n \omega_m t\right] \cos \omega_c t \tag{3-74}$$

where: m = the modulation coefficient

ω_m = the modulation radian frequency

ω_c = the carrier radian frequency

The b's in the expansion are determined by the curve-fitting process. If this expression is expanded into all of its frequency components, not only do the carrier frequency, ω_c, and the desired side-bands, $\omega_c \pm \omega_m$, appear, but spurious side bands at $\omega_c \pm n\omega_m$, $n \neq 1$, also appear. This method can be used to predict accurately the amplitudes of the splatter side-bands of typical plate-modulated transmitters; it requires an accurate curve of rf output versus plate voltage.

e. Transmitter Noise Emission. Transmitter noise (including hum, vibration, thermal noise) is that modulation of the carrier that is caused by noise generated in the various rf stages, together with noise from the audio system and power supply. The type of interference produced by transmitter noise does not have a characteristic audio output in a receiver being interfered with and is, therefore, not easily recognized. The output of any oscillator consists of discrete and continuous spectrums. The discrete spectrum comprises the desired frequency and its harmonics; the continuous spectrum is noise that results from such areas as oscillator frequency jitter, power-supply noise, and tube noise. Interference from the continuous noise spectrum is primarily adjacent-channel interference; it may, however, extend over a considerable number of channels for a high-power transmitter. In general, this noise does not degrade the desired signal appreciably because the depth of modulation or deviation due to noise is small when compared to the desired modulation. The amplitude of noise side-bands decreases rapidly with frequency separation from the carrier; however, it is possible for portions of the noise side-bands to extend into adjacent channels and be of sufficient magnitude to cause interference. Class C amplifiers tend to produce less noise output than linear amplifiers, and transmitter noise increases as the transmitter frequency increases.

f. Transmitter Intermodulation.

(1) Transmitter intermodulation emission results from the intermodulation products formed by mixing two or more components of a complex wave in a nonlinear stage within a transmitter. Intermodulation products are significant because they effectively extend the spectrum of out-of-band emission by the transmitter, thus increasing the likelihood of off-channel interference.

(2) A prerequisite for rf intermodulation is the presence of two signals and a nonlinear circuit element. If two or more signals are impressed on a nonlinear element, mixing occurs in such a way that all the harmonics and all of the sums and differences of all the combinations become potential signals. In the case of communications transmitters, the antennas may receive interfering signals that are conducted into the transmitters through the transmission lines. Once such an interfering signal reaches a nonlinear element such as a power amplifier, it can cause the generation of a broad band of inter-

modulation products and its own harmonics. The magnitudes of the resulting signals are dependent upon several factors, including the impedance of the circuits they encounter, the characteristics of the nonlinear elements, the relative strengths of the mixing signals, and the rf bandpass characteristics of the tuned circuit(s), transmission line and antenna involved.

(3) There are two types of transmitter intermodulation: audio and radio frequency. Audio intermodulation occurs in the modulator and rf stages, and may produce extraneous components in the passband and adjacent channels. Radio-frequency intermodulation requires the presence of an undesired signal, either internal or external to the transmitter, which may produce radiation at the undesired frequency, plus and minus multiples of the frequency spacing between the desired and undesired frequencies.

(4) The final amplifier stage of a transmitter is usually biased to operate in a nonlinear fashion, making it an efficient generator of intermodulation products. The mathematics of the generation of intermodulation products in a transmitter are very similar to those of receiver intermodulation; however, transmitter intermodulation is produced primarily by signals that enter the plate rather than those that enter the grid circuit. Only one external signal is required for transmitter intermodulation product generation. The frequency of such a signal, which differs from the desired signal frequency by only a few hundred kc, may be passed by the transmitter output circuitry at an amplitude sufficient to produce serious intermodulation products when mixed with the desired signal fundamental or its harmonics.

(5) Intermodulation products of three orders may be generated within a transmitter output stage: the first, primary mix intermodulation, results from mixing of the harmonics of a desired signal and an interfering signal; the second results from mixing of the fundamental desired signal with harmonics of the

interfering signal within the desired signal transmitter; and the third occurs when harmonics of the desired signal and harmonics of the interfering signal -- both of which are generated in the output stage of the desired signal transmitter -- mix. All three orders can be identified by:

$$f_s = f_d + \left(\frac{N+1}{2}\right) \Delta f, \text{ or } f_s = f_d - \left(\frac{N+1}{2}\right) \Delta f \qquad (3\text{-}75)$$

where: f_s = intermodulation product frequency

f_d = desired or carrier frequency

f_1 = interfering signal frequency

N = product order

$\Delta f = f_1 - f_d$

These equations yield two products for each order; one that is greater in frequency than the desired signal, and one that is lower in frequency.

(6) Intermodulation products, generated in an rf amplifier, are related to the plate current, which can be represented by a power series of the form:

$$i = a_0 + a_1 e + a_2 e^2 + a_3 e^3 \ldots\ldots \qquad (3\text{-}76)$$

The constants are determined by the tube characteristics and operating point, which, in turn, are dependent upon the circuit parameters. Because of variations in tube characteristics and circuit parameters, the constants must be evaluated for each particular circuit. If the rf amplifier components were perfectly linear, the 3rd, 4th, and higher-order coefficients would be zero; however, components, such as vacuum tubes, are not linear and have higher-order coefficients of significant magnitude. If we assume an input voltage to the rf stage of the form:

$$e = E_1 \sin \omega_1 t + E_2 \sin \omega_2 t, \qquad (3\text{-}77)$$

then the power series can be expanded to obtain the intermodulation products generated. The output, resulting from the linear term of the power series, is of the same form and frequency as the input; while the squared term results in a dc component, components at twice the input frequencies, and components at the sum and difference of the input frequencies. These products will not usually be of importance when considering rf intermodulation in the output stage of a transmitter. The output circuit is usually selective enough so that frequencies, far removed from the transmitter-tuned frequency, are greatly attenuated. The squared term, though not important in the generation of transmitter intermodulation, may produce serious interference in a receiver.

(7) The expansion of the cubic term produces a large number of terms with components at the input frequencies, f_1 and f_2, and at frequencies $3f_1$, $3f_2$, $2f_1 \pm f_2$, $2f_2 \pm f_1$. The intermodulation products of interest are those that are in the transmitter output passband. In general, they are the difference products, $mf_1 - nf_2$, where m and n are integers. The sum frequencies produce components that are considerably removed from the transmitter passband. The sum of m and n, usually referred to as the intermodulation product order, is the same as the exponent of the lowest power term in the power series that produces this particular frequency term. The amplitudes of these signals are proportional to $E_1^m E_2^n$. The harmonics of the carrier, necessary for the production of these intermodulation products, usually exist in the transmitter final amplifier at relatively high levels. The amplitude of the intermodulation products will be proportional to the n:th power of the amplitude of the signal that generates the intermodulation product:

$$E_s \propto E_1^n \qquad (3\text{-}78)$$

where: E_s = intermodulation product amplitude

E_I = interfering signal amplitude

n = order of the interfering signal required to generate intermodulation product of order m + n

g. Shock-Excitation Emission. Shock-excitation emission is transmitter radiation that results from external rf excitation in linear elements of a transmitter. When a short-duration pulse signal is applied to a resonant circuit, the circuit oscillates for a number of cycles. The oscillation assumes a damped-wave pattern; the number of cycles determined by the Q of the resonant circuit, and the frequency of oscillation determined by the natural resonant frequency of the circuit. Upon reaching rf or if resonant circuits, short-duration, high-amplitude pulses cause the resonant circuits to generate damped waves at the natural resonant frequency. Spurious outputs can also result from shock excitation at frequencies other than the fundamental frequencies of nearby transmitters.

h. Transmitter-Coupling Emission.

(1) Transmitter-coupling emission in a transmitter results when intermodulation products are generated as a result of the mixing of one or more signals in the transmitter. These signals are interference signals from other transmitters. When the antennas or transmission lines of two transmitters are located near each other, an appreciable rf voltage from one transmitter may be impressed across the output tank-circuit of the other. Because of nonlinear phenomena in the final amplifier circuit, this induced voltage can cause the generation and radiation of spurious frequencies at other than the operating-carrier frequency of either transmitter.

(2) The spurious output, f_s, of a transmitter with carrier f_1, when coupled to a transmitter with carrier f_2, may be obtained from the following:

$$f_s = \left| n_1 f_1 \pm n_2 f_2 \right| \qquad (3\text{-}79)$$

This equation yields the intermodulation products of all orders; third-order products are particularly troublesome in vhf systems.

(3) Interference-level prediction for third-order products requires a knowledge of various attenuation quantities existing between the desired transmitter and receivers and the undesired transmitter. The attenuations involved are shown on figure 3-156. With $A(2f_1 - f_2)$ as the attenuation of the third-order product with respect to the carrier level of transmitter number T_1 (if there were no path losses between T_1 and T_2), then A_T is defined as:

$$A_T = A_1 + A_1^1 + A_2 + A_3 + A_4 + A_{12} + A_{13} + A(2f_1 - f_2) \qquad (3\text{-}80)$$

Interference level (IL) is then obtained from:

$$IL = J = A_T \quad \text{(in db)} \qquad (3\text{-}81)$$

where: A_1 = attenuation between transmitter T_1 and antenna, at frequency $(2f_1 - f_2)$, in db

A_1^1 = attenuation between transmitter T_1 and antenna, at frequency f_2, in db

A_2 = attenuation between transmitter T_2 and antenna, at frequency f_2, in db

A_3 = attenuation between receiver R_3 and antenna, at frequency $(2f_1 - f_2)$, in db

A_4 = db of rejection of receiver front end with respect to receiver tuned frequency

A_{12} = path attenuation between transmitting antennas at frequency f_2, in db

A_{13} = path attenuation between transmitter T_1 and receiver for third-order intermodulation products generated in the transmitter

W = the db value obtained from the ratio of the output voltage of the desired transmitter, e_1, to the receiver sensitivity, e_r, both at a specified impedance

then: $W = 20 \log \frac{e_1}{e_r}$, $e_1 = \sqrt{PR}$ (3-82)

where: P = output power

R = impedance

A positive value of the interference level indicates that the third-order product power at the receiver terminals is above the receiver threshold; therefore, the system experiences interference.

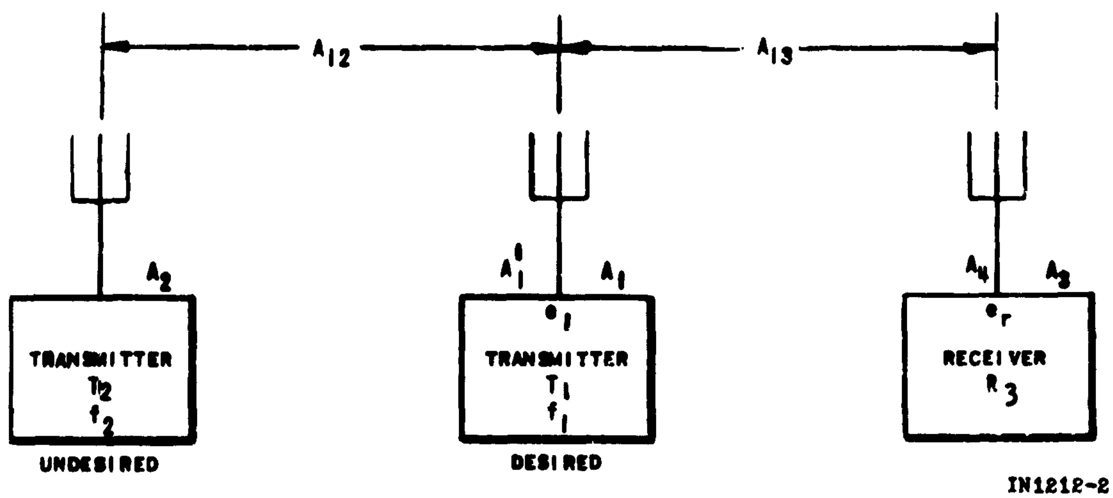


Figure 3-156. Transmitter Coupling Attenuation

3-43. Transmitter Interference Reduction

Transmitters cause interference by radiation and by conduction. The interference may be in the form of spurious radiation, broadband or narrowband emanations, modulation splatter, or intermodulation. Harmful interference may be caused by a single transmitter's emissions, or by the contributions from hundreds of transmitters. These causes of interference can be classified as spectrum congestion and system incompatibility. In preventing or minimizing transmitter interference, both the transmitting and receiving environments require attention. If, for a particular application, a frequency is causing trouble in adjacent equipment, the interference can be eliminated by traps or stubs; if, however, the transmitter is a production type that must be used in many different installations, suitable filters should be designed to suppress all the outputs to a negligible level. These filters should be placed in the power leads, antenna leads, and control leads. Also, to preclude case radiation, cases must be sealed by rf gaskets and perforated screens to prevent rf leakage. The maximum protection possible against interfering signals should be provided at all times.

a. <u>Frequency Generation.</u> Certain design practices that reduce spurious signal generation should be followed whenever applicable:

1) Tuned coupling networks should be used to prevent the passage of spurious signals to succeeding stages

2) Oscillator circuits should be decoupled from both input and output stages

3) The power amplifier should not be used as a frequency multiplier because it will contain a large undesired component at the fundamental or excitation frequency

4) The output stage should be designed for low distortion at the required power output to minimize the need for excessive grid excitation and heavy loading, both of which lead to the generation and transfer of spurious signals into the antenna

5) The use of frequency multiplication to generate the carrier frequency should be avoided except when absolutely necessary

6) Where applicable, multituned devices should be used for all tuned coupling circuits

7) The frequency stability of a transmitter should not be less than that required to ensure proper operation of the associated receivers. Improved transmitter stability results in reduced receiver bandwidth requirements. Many receivers are intentionally designed

with broad selectivity characteristics, sometimes at the expense of sensitivity, to compensate for transmitter drift. Ultimately, stabilities of the order obtainable by crystal control should be maintained

If these measures are followed, together with careful shielding and decoupling of the individual stages, the spurious content of the rf carrier can be reduced to a low level.

(1) Mixers are sometimes used in the frequency-generating scheme of transmitters. In some transmitter-receivers, the receiver local oscillator is also used for the transmitter. A signal from a crystal oscillator is added to the signal from the receiver local oscillator in a mixer to produce the transmitter frequency. Spurious emissions are generated in the mixer. For example, when a signal from a crystal oscillator at frequency, f_1, is added in a mixer to a signal at frequency, f_2, to produce a desired signal of frequency, $f_1 + f_2$, then signals are also produced at such frequencies as $f_1 - f_2$, $2f_1 \pm f_2$, $f_1 \pm 2f_2$ and $2f_2 \pm 2f_1$, etc. One way to eliminate spurious radiations, originating in a mixer, is to eliminate the mixer and use a direct means of frequency generation. If, however, a mixer is used in generating the transmitter frequency, the spurious signals generated there should be attenuated to sufficiently low levels by the tuned circuits following the mixer. A double-tuned circuit in the mixer output circuit and a double-tuned circuit in the driver-plate circuit will usually be satisfactory in reducing the level of spurious signals. Good shielding should be used between stages, and rf paths should be decoupled to keep the spurious signals from reaching the power amplifiers. Spurious radiations, whose frequencies are harmonics of the carrier frequency and which originate in the power amplifier, can be reduced by a double-tuned circuit. A typical circuit is the inductively-coupled antenna tuning circuit shown on figure 3-157.

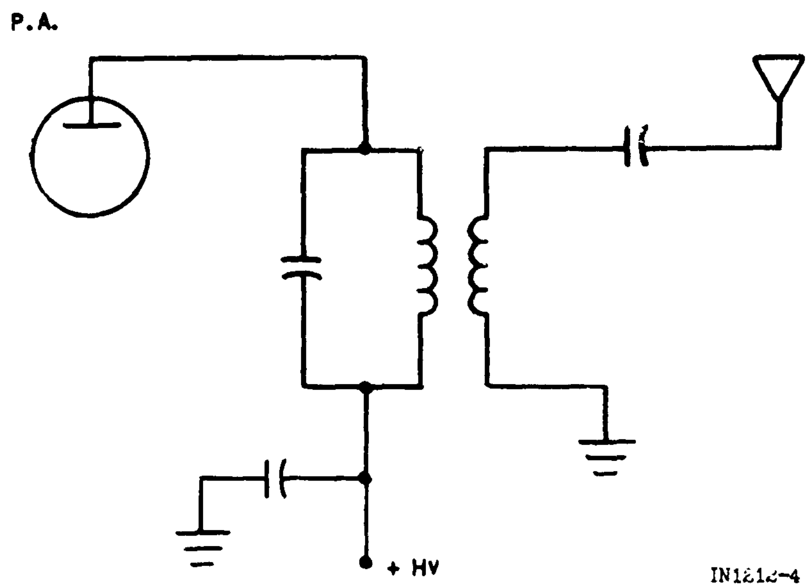


Figure 3-157. Double-Tuned Tank Circuit

(2) Shunt feeding of the high-voltage supply to the final stage plate is employed in some transmitters; and a tap on the plate tank is used to couple to the antenna. This type of coupling should be avoided because it eliminates the additional tuned circuit of the antenna coupling and tuning circuit. A pi-section tank circuit is used in some transmitters because of the wide variation in antenna impedances through which the transmitter must work. A pi-section, although similar to a low-pass filter section, is usually ineffective in suppressing harmonic radiation. A double-tuned tank circuit should be used in preference to a pi-section tank circuit. A double-tuned tank circuit will reduce harmonics more effectively than a single-tuned one.

(3) A comparison of the 2nd, 3rd, 4th, and 5th harmonics of three typical transmitters is shown in table 3-9. In this table, the transmitters are labelled as to type of output tank circuit. The transmitter with the double-tuned tank has its highest spurious radiation level 45 db below the carrier, while the single and pi-section transmitters are 10 and 9 db higher, respectively.

The levels shown in the table are typical levels to be expected in transmitters with no extra circuitry to reduce harmonic radiation. To reduce the harmonic radiation further, low-pass filters must be incorporated between the tank circuit and the antenna. A simple filter, consisting of a constant-k, pi mid-section and two m-derived terminating half-sections, will give approximately 40-db attenuation to most of the harmonics, depending upon the carrier frequency. More elaborate low-pass filters can be built to suppress harmonics by 60 db. For transmitters operating above approximately 100 mc, re-entrant cavity filters can be used. These have the added advantage of being band-pass rather than low-pass filters. A combination of a double-tuned output circuit and a 60-db low-pass filter will reduce harmonic radiation approximately 105 db. Good shielding of the filter is necessary to prevent radiated coupling from bypassing the filter.

TABLE 3-9. LEVEL OF HARMONICS BELOW CARRIER FOR THREE TYPES OF TANK CIRCUITS

Harmonic	Pi-Section	Single-Tuned	Double-Tuned
2	36 db	35 db	45 db
3	47 db	61 db	64 db
4	54 db	61 db	65 db
5	60 db	67 db	73 db

(4) The use of a low-pass filter introduces another problem in transmitter design. Many transmitters operate over frequency ranges that cover one or more octaves. It is very difficult for one filter to attenuate the second harmonic of a signal at the lower end of an octave and pass a signal at the high end of an octave. Therefore, it is necessary to use more than one filter to cover the octave range. The problem is further compounded in transmitters that switch tuning ranges in octave bands. For example, in a transmitter that has an exciter which has four bands (1.5 to 3, 3 to 6, 6 to 12, and 12 to 20 mc), the first three bands are octave bands. If low-pass filters are used in transmitters with

such an arrangement, two low-pass filters would have to be used in each octave band, and provisions for switching filters somewhere within each band would have to be made. Another technique would be to design a multiband transmitter with each band covering less than one octave, and then switch in a separate filter for each band.

b. Channel-Width Considerations. A basic difference between desired and interfering signals is their frequency content: an interfering signal can be spread over a considerable portion of the frequency spectrum; a desired signal usually lies within a well-defined region of the spectrum. One of the most important considerations in the design of an interference-free communications system is the channel width. At the present time, fm, am, and single side-band systems are being operated simultaneously in the same frequency bands, and within channel separation distances that permit the systems to interfere with each other. It is necessary, therefore, to choose channel widths that will accommodate all types of modulation; otherwise there will be adjacent channel interference between systems. If a channel is too wide, valuable frequency spectrum will be partially wasted. If a channel is too narrow, stringent requirements are placed on the equipment, usually at added cost, and increased interference may result. A communications system's operational channel width should be chosen on the basis of the following factors:

1) Accuracy and stability of the receiver

2) Accuracy and stability of the transmitter

3) Frequency changes due to Doppler shift

4) Modulation bandwidth

5) Guard-band requirements

Accuracy and stability considerations are of prime importance. If a transmitter drifts sufficiently far from the center of its channel, modulation sidebands can splatter into an adjacent channel. If a receiver drifts sufficiently far from the desired center of its channel, it can receive modulation interference from the adjacent channel, which can cause cross-modulation with resulting high-frequency chatter in the receiver. If a transmitter is sufficiently detuned relative to a receiver that is receiving its signal, severe loss of intelligibility and operating distance can result. Transmitter bandwidth should be held to a minimum consistent with overall circuit requirements.

c. Component Selection. Components utilized in rf circuitry should be designed for frequency ranges beyond those in which the equipment is intended to operate. In the selection of a coil for use in a tuned circuit, for example, the impedance of the coil is a function of frequency. The distributed impedance of the coil can cause multiple resonances: there will be as many resonances as there are frequencies at which the inductive reactance is equal to the capacitive reactance. The solution to this problem is to select only components whose inductive reactance and capacitive reactance produce resonances at frequencies greater than the operating frequency. In general, this means that high-quality components should be used, lead lengths should be kept short, and good judgement should be exercised in component placement.

d. Oscillator Stages.

(1) Communications transmitters are rarely operated at frequencies such that the desired rf output and oscillator frequencies are the same. In many transmitters, the output frequency is a multiple of a basic oscillator frequency. This means that, unless proper precautions are taken in the design of the transmitter, each harmonic of the oscillator frequency may appear on the transmitter antenna. There are several ways in which spurious oscillator outputs can be controlled and contained. Complete enclosure of the oscillator for the purpose of shielding and containing the radiation is general practice in well-built transmitters. Metal walls and metallized gaskets for all covers on an enclosure should be used to restrict the radiation of rf energy from the oscillator. All nonsignal circuits should be filtered.

(2) Most transmitters use a common power supply for several low-power stages and the basic oscillator. There is no objection to this practice provided proper precautions are taken to minimize the undesired transfer of rf energy between the stages. Common filament or heater leads can be serious offenders in many transmitters of this type. Where a single transformer is used to supply heater voltage to several stages, one or more of the following measures will be found of assistance in interference control:

1) Bypass all heater leads to ground with mica capacitors at each electron tube socket, using the shortest possible capacitor leads

2) Insert rf chokes of adequate current-carrying capacity in heater leads at the electron tube sockets, and bypass as in 1)

3) In extreme cases, use series, shunt, pi, or multiple-tuned networks

4) Bypass all power and control leads entering the oscillator enclosure at their points of entry

(3) In transmitters where the basic oscillator frequency is multiplied at a low level and then amplified, spurious radiations, at every frequency present in the low-level stages, invariably appear in the transmitter output. Signals, at harmonics of these frequencies, also often appear. These radiations should be reduced to acceptable levels. To do this, use as few stages of frequency multiplication as possible. In transmitters that employ self-controlled oscillators, only two stages of frequency multiplication should be used. It is best to use the oscillator frequency straight through. The tuned circuits of the multiplier and driver stages in the transmitter have to reject signals at frequencies lower than that to which they are tuned. The greater the separation of these frequencies, the greater the rejection; the lower the order of multiplication, the greater the separation of frequencies.

Furthermore, the lower the order of multiplication, the fewer the number of spurious frequencies generated. For example, an am transmitter, operating in the 116- to 132-mc band, has a frequency multiplication of eighteen. The crystal oscillator operates near 7 mc and is followed by two triplers and a doubler. Spurious radiations existed at every multiple of 7 mc up to 1000 mc. By changing the oscillator to 14 mc and multiplying the frequency 9 times, one-half of the spurious radiations were completely eliminated.

(4) In transmitters in which it is necessary to employ frequency multiplication, double-tuned circuits should be used between all stages. In addition, there should be at least four tuned circuits, tuned to the output frequency, ahead of the final amplifier. A double-tuned circuit between the last multiplier and the driver, and a double-tuned circuit between the driver and the final amplifier, can reduce the spurious radiations generated in the low-level stages more than 80 db below the carrier. Good shielding between rf circuits is necessary to prevent coupling of undesired signals from stage to stage. If these measures are taken when designing transmitters, spurious radiation can be kept to levels of 80 db, and more, below the carrier.

e. Output and Antenna Stages. When designing the output stages and antenna circuit, several design procedures should be kept in mind:

1) It should be possible to obtain the nominal output over the entire tuning range

2) The operating angle of power output tubes should be selected to reduce the number and magnitude of spurious signals

3) Push-pull output stages should be used

4) The antenna circuit should be filtered with either a low-pass or a band-pass filter

5) Nonlinear elements should not be used in the antenna circuit

6) The antenna-feed circuit should be completely shielded

7) Antenna gain and front-to-back ratio should be as high as practicable

8) Antenna beamwidth in azimuth should be the minimum practicable

9) Antenna beamwidth in elevation should be the minimum practicable

10) Antenna sidelobes should be minimized

f. Power Output. In general, transmitters should have means for varying radiated power continuously, so that, in the interest of reducing interference, only the minimum power required for satisfactory operation is radiated.

g. Output Filters.

(1) Many techniques are possible to minimize spurious emissions and responses; use of filters for transmitters and receivers represents the direct approach. Most filters have periodic characteristics; that is, a filter will attenuate certain prescribed frequency bands, but will pass many other bands spaced periodically throughout the spectrum.

(2) Waveguide filters, that are capable of handling peak power values of the order of 5 megawatts with insertion losses of only 0.1 db, can be fabricated for harmonic suppression. The use of such filters can alleviate spurious emission conflicts in receiving systems whose operating frequencies are considerably removed from the frequency band of the transmitter in question. A tunable filter with a narrow bandwidth can reduce, if not eliminate, interference between transmitters using the same frequency band.

(3) Harmonic-suppression filters can be used at the output of transmitters, before undesired frequencies reach the antenna, or in the antenna input circuits of receivers. Ideally, the filters should be installed as close as physically possible to the interference sources. Some of the factors to be considered in

the proper use of a harmonic filter are:

1) In the pass-band, the filter should perform as a conventional transmission line section, and the filter should not cause an insertion loss greater than 0.5 db, or a VSWR greater than 1.1. If there is a mismatch, it should not reduce the output power of the transmitter below a specified level

2) In the stop-band, the attenuation should be at least [illegible] db, or enough to bring the spurious emissions to the level required by the applicable specifications, whichever is greater

(4) Wherever good engineering practice permits, harmonic filters should be the low-pass type because frequencies below the fundamental are usually at a lower power level than the harmonics. In cases where the fundamental is formed by frequency synthesis, and strong subharmonics are present, a band-pass filter should be used. The preferred cut-off frequency of a harmonic filter depends on the individual application. For example, if a transmitter has a desired fundamental frequency, f_o, of 100 mc, and the first spurious frequency (the second harmonic) is $2f_o$, or 200 mc, a good choice for the cutoff frequency is 1.5 f_o, or 150 mc.

(5) Conventional filters, consisting of a combination of constant-k and m-derived sections using lumped parameters, can be used successfully up to 100 mc. However, as the frequency is increased, it is increasingly more difficult to develop mass-produced units using lumped parameters. For this reason, at the higher frequencies, transmission-line filter elements or waveguide filters can be used.

h. Spurious and Harmonic Emission.

(1) Reduction of spurious outputs in transmitters should be considered in the initial design of the oscillator. Elimination of spurious outputs not only prevents interference, but in-

creases the fundamental power output of the transmitter. Unless a stage is absolutely linear in operation, spurious or harmonic frequencies will appear in the plate circuit. In the Class C power amplifier of a transmitter, considerable harmonic current exists. These harmonics appear at the output unless they are eliminated by tuned circuits; consequently, the selectivity of the tuned circuits following the final power amplifier determines the ultimate harmonic content of the transmitter output. The frequency of the spurious and harmonic emissions from a transmitter using a master oscillator can be identified by:

$$f_s = \frac{nf_o}{k} \tag{3-83}$$

Those spurious and harmonic emissions, from a transmitter utilizing frequency synthesis or automatic frequency control, may be identified by:

$$f_s = \left| \frac{n\,f_{lo} \pm m\,f_{to}}{p} \right| \tag{3-84}$$

where: f_s = frequency of spurious emission

m,n = integers 0, 1, 2, 3, ...

p = 1, 2, 3, ...

f_o = transmitter operating frequency

f_{mo} = transmitter master oscillator frequency

$k = \frac{f_o}{f_{mo}}$ = frequency multiplication factor

f_{lo} = local oscillator frequency

f_{to} = transmitter oscillator frequency

(2) Usually, the envelope of the spurious and harmonic emissions does not decrease linearly with frequency; rather, it tends to undulate. In transmitters of the same type, variation of para-

meters such as wiring, circuit component values, biasing levels, and vacuum tube characteristics will shift the frequency spectrum of the envelope. This displacement results in variation in the levels of harmonics from one transmitter to another. There is no definite cutoff point, with respect to frequency, for either the harmonic or spurious emissions.

(3) Tubes should be operated in a mode that generates a minimum number of harmonics. Design engineers can use Fourier analysis of the waveform to predict the harmonic content of a particular output. Such an analysis for the output of a Class C amplifier, is presented: this type of an analysis can be applied to any type of circuit. Assume that the signal, e_g, of a tube-operated Class C amplifier, is composed of a number of harmonic sinusoidal signals, and that the lowest-frequency signal present is at radian frequency ω_1, then:

$$I_p = \frac{A_o}{2} + A_1 \cos \omega_1 t + A_2 \cos 2\omega_1 t + \ldots + B_1 \sin \omega_1 t + B_2 \sin 2\omega_1 t + \ldots \tag{3-85}$$

where:

$$A_n = \frac{2}{T}\int_o^T I_p \cos n\omega_1 t \, dt$$

$$B_n = \frac{2}{T}\int_o^T I_p \sin n\omega_1 t \, dt$$

and:

$$T = \frac{2\pi}{\omega_1} = \frac{1}{f_1}$$

If the conduction angle of a tube is specified, and, if the amplitude of the plate current during conduction is specified as a function of the tube characteristics, plate-supply voltage, grid-bias voltage, and grid-driving voltage, it is possible to solve for the coefficients A_n and B_n for each value of n that is of interest. This method requires measured data for the conduction angle and plate impedance of the output tube, and for the spectrum of the grid-driving signal.

(4) To reduce or eliminate harmonics, filters can be placed in the output circuit of the transmitter. In addition, the use of suitable coupling circuits, such as filters or tuned traps, will reduce the generation of harmonics. Filters used should be capable of handling the fundamental power frequency with a minimum of insertion loss. Bypass filters in both B+ and filament circuits will prevent harmonics from reradiating to other circuits.

(5) Guy-wires should be divided into insulated sections, the lengths of which should not be harmonically related to the operating frequency. Such a division prevents the excitation and reradiation of spurious transmitter radiations by resonant lengths of guy-wire. Abandoned antennas and guy-wires should be completely removed. Merely grounding an unused piece of wire is not sufficient to eliminate it as an interference source.

(6) All joints and connections exposed to the weather should be designed for ease of inspection because oxides formed at joint interfaces may act as nonlinear elements and cause the generation of spurious signals. Antenna and ground connections should always be made so that an oxide film cannot form between contacting parts. Joints should be waterproofed, if possible.

(7) Provisions should be made to allow transmitters to be tuned for maximum attenuation of spurious radiation. Transmitters, using variable frequency oscillators, should be tuned to the operating frequency with a frequency meter, and checked periodically to ensure on-frequency operation. Crystals in transmitters should be

checked periodically to ensure that the transmitters are operating at the correct frequency. Output power and bandwidths should be maintained at the minimum level consistent with reliable communications. Radar output spectra should be monitored to determine whether excessive sidebands are present, and whether the magnetron requires frequency adjustment or replacement.

(8) Spurious external cross-modulation interference may be generated by a metallic device exposed to strong electromagnetic fields. Such interference will be generated when such device:

a. contains a self-resonant circuit at

$$f_r = nf_1 \pm Nf_2 \qquad (3\text{-}86)$$

where: f_r is the self-resonant frequency of the device

f_1 and f_2 are the frequencies of the electromagnetic fields

n and N are integers

and:

b. when the device contains a non-linear circuit element coupled to the device self-resonance.

Metal building framework, gutters and down-spouts, lightning rods and ground leads, fences, open-wire lines, stoves and stove pipes, metal doors and door tracks, are all examples of metallic devices capable of creating such interference. All such devices, required to be located within high electromagnetic-field ambients, should be designed with suitable bonds and grounds, and be adequately maintained to prevent formation of metal/metal oxide nonlinear interfaces between elements. In some instances, "spoiler" elements may be added to the device to detune structural resonance.

i. Parasitic Emission.

(1) Within a transmitter, parasitic emissions can arise from such sources as interconnecting leads between tubes, and components

that may develop significant power at their own self-resonant frequency. Shock excitation of various resonant circuits, by transient signals within the transmitter, can also produce parasitic emission. These emissions may emanate from oscillators, amplifiers, and almost any other electronic circuit. In addition to radio interference, these spurious frequencies may create distortion in linear amplifiers and modulators; causing undesired side-bands in transmitter outputs, flashover, reduction of useful power output, and other undesirable effects. Parasitic emissions are difficult to locate and, in general, cannot be measured separately from the overall transmitter emission pattern. Procedures, for identifying this type of emission, should be worked out during spurious-emission measurements.

(2) Methods of suppressing parasitic oscillations are basically the same as for spurious emissions. Oscillations may frequently be suppressed by minor circuit modifications, such as the insertion of a small resistor, or the replacement of a choke coil. The first step in suppression of these oscillations is to locate the offending circuit. When it has been located, necessary modifications should be made; these should not affect the normal operation of the circuit. Radio-frequency amplifiers can be neutralized by feedback circuits, or by screen-grid neutralization. A small resistor, of from 1.0 to 25.0 ohms, can be inserted in series with the grid or plate lead of the tube in the offending circuit; it should be sufficient to damp out the parasitic oscillations. If the cause of the oscillations is a stray resonance between an rf choke and its own distributed capacitance, or some other capacitance in the circuit, another choke with different inductance can be substituted. The insertion of a small resistor, in series with the choke, will also eliminate the parasitics. Sometimes, a slight detuning of certain elements, not enough to cause any detrimental effect to normal operation, is effective in eliminating parasitic oscillations.

j. Side-band Splatter. Suppression of side-band splatter is accomplished by increased filtering, improved linearity, rf feedback, and regulation of the power supply. Nonlinear modulation limiters produce a considerable amount of side-band splatter. Modulators should use linear amplifiers, where possible, and have their outputs filtered to generate only the desired frequency spectrum.

k. Modulation Splatter.

(1) Interference caused by modulation splatter can be effectively controlled by audio processing and modulation limiting. Speech processing and modulation-limiting methods vary with the type of modulation.

(2) Splatter may be minimized, in transmitters employing voice-band channels, by use of limiting amplifiers to avoid overmodulation. This approach is effective on transmitters using amplitude, frequency, phase, and single or double-single side-band modulation schemes. When multiple voice-band channels are used, such as two 3-kc channels on both the upper and lower side-bands of a double single-side band transmitter, limiting amplifiers should be used on each voice-band input. Each input should contain a suitable band-pass filter, limiting the frequency range to 300-3000 cycles.

(3) Peak clipping may be employed to remove the large (6-9 db) transient spikes present in the audio spectrum of the average male voice. Waveform clipping introduces a large amount of odd-order harmonic distortion. The amount of clipping must be restricted to avoid generation of side-band components which might cause more interference than the unclipped wave forms. Generation of odd-order harmonic distortion may be avoided by:

1) Separating the voice-band spectrum into third-octave segments in suitable filters

2) Clipping each segment

3) Removing the odd-order harmonics by passing each spectral segment through a second third-octave filter

4) Reconstituting the voice-band spectrum by combining the third-octave segments in a suitable amplifier

(4) The absolute power levels of the sidebands of an fm transmitter may be computed from the following:

$$W_p = W_o - 20 \log \left[\frac{1}{J_p\left(\frac{\Delta F}{f_m}\right)} \right] \quad \text{dbw} \qquad (3\text{-}87)$$

where: W_p = the absolute power level of the sidebands

W_o = the unmodulated carrier power in decibels below one watt

$J_p\left(\frac{\Delta F}{f_m}\right)$ = Bessel coefficient of the first kind of order p and argument $\left(\frac{\Delta F}{f_m}\right)$

ΔF = frequency deviation

f_m = modulation frequency

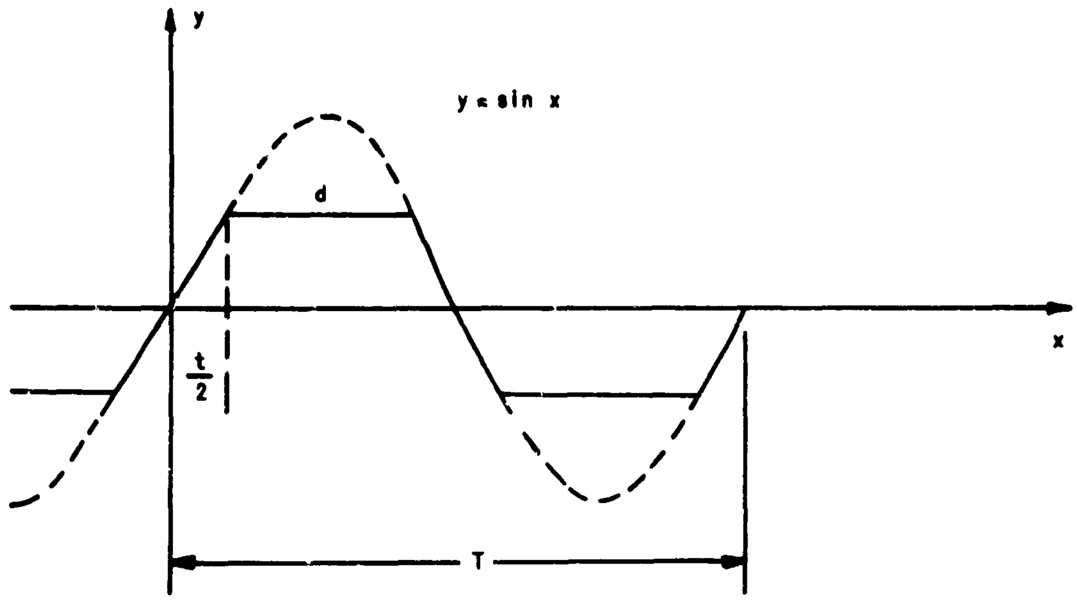
(5) A severely clipped sinusoidal tone (fig. 3-158A) may be closely approximated with a symmetrical trapezoid wave (fig. 3-158B).

The nth spectral-component amplitude of such a wave is given by:

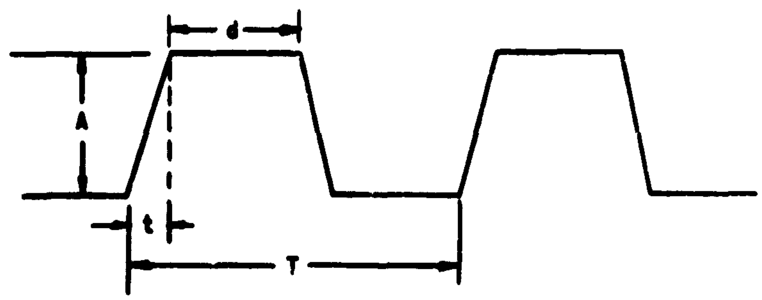
$$C_n = \frac{2A(t+d)}{T} \left[\frac{\sin(n\pi t/T)}{(n\pi t/T)} \right] \cdot \left[\frac{\sin\left[n\pi(t+d)/T\right]}{n\pi(t+d)/T} \right] \qquad (3\text{-}88)$$

If the degree of clipping of positive and negative half-cycles is equal, then $(t + d) = T/2$ (fig. 3-158B) and:

$$C_n = A \left[\frac{\sin(n\pi t/T)}{n\pi t/T} \right] \cdot \left[\frac{\sin(n\pi/2)}{n\pi/2} \right] \qquad (3\text{-}89)$$



A. CLIPPED SINUSOIDAL TONE



A = PULSE AMPLITUDE
t = PULSE BUILD-UP TIME AND DECAY TIME
d = PULSE WIDTH AT MAXIMUM AMPLITUDE
T = PERIOD

B. TRAPEZOIDAL WAVEFORM IN1212-3

Figure 3-158. Approximation of Clipped Sinusoidal Tone by Trapezoid Waveform

The value of t is found from a specification of the amount of clipping, N, in db:

$$N = 20 \log \left[\sin \left(\frac{2\pi}{T} \cdot \frac{t}{2}\right)\right] \qquad (3\text{-}90)$$

$$t = \frac{T}{\pi} \sin^{-1} \left[\text{antilog } \left(\frac{N}{20}\right)\right] \qquad (3\text{-}91)$$

As an example, consider the case for 10 db of clipping (N = - 10 db):

$$t = \frac{T}{\pi} \sin^{-1} \left[\text{antilog } -0.5\right]$$

$$t = .102\ T$$

The amplitude of the nth harmonic is then:

$$C_n = A \left[\frac{\sin\ (.102\ n\pi)}{(.102\ n\pi)}\right] \left[\frac{\sin\ n\pi/2}{n\pi/2}\right]$$

A simplified expression for the worst-case amplitude of the nth harmonic, ignoring sign, can be obtained by letting

$\sin\ (.102\ n\pi) = 1$ and $\sin\ (n\pi/2) = 1$

Then, for $n \geq 5$ and odd:

$$C_n \approx \frac{A}{(.102\ n\pi)\ (n\pi/2)}$$

$$C_n \approx 1.99 \frac{A}{N^2}$$

Using this expression for harmonic amplitude, the powers radiated in the sidebands due to distortion products can be calculated as follows:

The general expression for a Bessel coefficient of the first kind is:

$$J_p\left(\frac{\Delta F}{f_m}\right) = \sum_{k=0}^{\infty} \frac{(-1)^k}{k!\ (p+k)\ !} \left(\frac{\Delta F}{2f_m}\right)^{p+2k} \qquad (3\text{-}92)$$

If $(\Delta F/f_m)$ is small, the first term of the series is a reasonable approximation.

$$J_p\left(\frac{\Delta F}{f_m}\right) \approx \frac{1}{p!}\left(\frac{\Delta F}{2f_m}\right)^p ; \quad \frac{\Delta F}{f_m} \leq 0.1, k=0$$

Since the amplitude of the sideband components, $J_p(\frac{\Delta F}{f_m})$, diminishes very rapidly for small $(\frac{\Delta F}{f_m})$, only the case $(p = 1)$ need be considered:

$$J_1\left(\frac{\Delta F}{f_m}\right) \cong \frac{1}{2} \cdot \frac{\Delta F}{f_m} \qquad (3\text{-}93)$$

Thus, the relative amplitude of the first order sideband is $\frac{1}{2}$ $\frac{\Delta F}{f_m}$, and the related relative power is $20 \log \frac{1}{2} \frac{\Delta F}{f_m}$.

l. Transmitter Noise Emissions. Reduction of transmitter noise can best be accomplished in the design stage by providing suitable filters following the audio stages, mixers, modulators, and oscillators. An optimum design including these filters will have a noise output that is determined, to a major extent, by characteristics of the final amplifier stage.

m. Intermodulation. Several techniques may be employed to reduce the effects of transmitter intermodulation. Frequency assignments should be established, where practicable, so that the third- and forth-order products do not fall on, or immediately adjacent to, desired receiver frequencies. Selective filters can be used to provide additional selectivity (for example, isolation) between transmitters. Linear-amplifier techniques can be used to reduce the basic intermodulation product that is generated. The generation of the carrier frequency by a mixing process should be avoided. Only when absolutely necessary should frequency multiplication be used to generate the carrier frequency. All tuned coupling circuits should be multituned devices.

n. Blanking. Intensity-modulated displays, that must operate in a pulsed-interference environment, lend themselves particularly to the use of blanking techniques. By blanking the display during periods of transmission by an offending transmitter, and causing the blanked region to move constantly on the display, virtually all interference can be removed with very little loss of useful information. Where a number of transmitters and receivers must be located in the same vicinity, blanking can be accomplished by direct interconnection. Another approach for this application involves the use of a self-contained blanker. The blanker consists of a wide bandpass gating receiver containing delay circuitry and placed ahead of the normal system receiver. The gating receiver is set to blank the system whenever the spectrum components of a transmitter appear above an objectionable level.

o. Signal Integration. A number of integration techniques can be employed effectively to reduce interference in such equipment as radar units. They make use of the fact that, while a target signal appears at approximately the same range on a number of successive returns, random interfering pulses do not. If several successive returns are required to exceed a preset threshold level, the random pulses never appear at the output. Radar tracking circuitry usually includes a simple form of integration. The characteristics of phosphors, used in conventional intensity-modulated cathode-ray-tube displays, do not provide sufficient interference rejection by integration. Storage tube devices, on the other hand, are particularly well suited to this application because of their controllable integration characteristics.

p. Effects of Transmitter Spurious Outputs. To predict the effects of spurious signals on nearby receivers, the following computation may be used:

$$P_R = (P_T - K_S - K_T + G_T) - 10 \log 70 \frac{d^2}{\lambda_s^2} + (G_R - K_R) \tag{3-94}$$

where: P_R = power at the receiver antenna terminals (dbw)

P_T = power transmitted at the fundamental frequency (dbw)

λ_s = wavelength of fundamental, harmonic, or spurious signal (meters)

K_S = attenuation of signal (λ_s) below the fundamental signal (db)

K_T = attenuation of transmitting antenna feed system at (λ_s)

G_T = gain of transmitting antenna with respect to a dipole (db) at λ_s

d = transmitting-to-receiving antenna separation distance (meters)

G_R = gain of receiving antenna with respect to a dipole (db) at λ_s

K_R = antenuation of receiving antenna feed system at λ_s

All factors in the equation must be known to make accurate predictions. Such complete data, however, is not generally available, and certain approximations must be made. While P_T, K_S, d, and λ_s must be known, reasonable values can usually be assumed for K_T, G_T, G_R and K_R. All available data should be used to arrive at as close an approximation as possible. When the antenna feed system consists of a waveguide, the attenuation (K_T or K_R) below cutoff can be computed from:

$$K_T \text{ or } K_R = \frac{54.5\,\ell}{\lambda_c} \sqrt{1 - \left(\frac{\lambda_c}{\lambda_s}\right)^2} \text{ db} \qquad (3\text{-}95)$$

where: ℓ = length of antenna feed guide (meters)

λ_c = cutoff wavelength of waveguide (meters)

At frequencies above cutoff (for the relatively short lengths involved in antenna feed systems), it is safe to assume no attenuation; K_T or $K_R = 0$. For lower frequency systems that do not employ waveguide feeds, K_T and K_R can also be assumed equal to zero. It is frequently necessary to approximate the transmitting and receiving antenna gains as well: here again $G_T = G_R = 0$ can be used.

Section VII. ANTENNAS

3-44. General

a. Controlling the generation, propagation, and effects of interference from antennas, will enable the design engineer to minimize the transmission or reception of undesirable or spurious signals. Such signals are radiated by antennas in various frequency ranges, modes, signal intensities, and directions. A perfect antenna would transmit a signal only of the desired frequency, power, and mode; and in a direction determined by the design criteria of the antenna. In practice, however, the perfect antenna does not exist, and undesirable or spurious signals are radiated or received.

b. Interference problems arise when an antenna radiates undesirable energy from an energy source. An antenna, located in the electromagnetic field created by this undesirable energy, will collect the energy and transfer it into a receiver or other device to which it is connected The ease with which this occurs depends upon the particular characteristics of the antenna. The same antenna properties, that allow energy to propagate, can be used to impede its transfer. Antennas, that keep the transmittal of undesirable signals to a minimum, can be selected by following such design practices as increasing the antennas' directivity, reducing side-lobe sensitivity, or eliminating back lobes.

3-45. Types of Antennas

Antennas take many forms and sizes, such as dipoles, whips, parabolic reflectors, and horns. Some types of antennas, and their normal frequencies of use, are shown in table 3-10. The choice of antennas depends on their use and such other factors as:

1) Whether omnidirectional, directional, fixed, rotating, or tracking

2) Scan rate

TABLE 3-10. TYPES OF ANTENNAS

0.150-30 mc	30-60 mc	60-500 mc	500-20,000 mc	20-40 kmc
Long & short range communications	Long & short range communications	VHF communications	Point-to-point communications	
	Long-range radar & guidance	Long-range radar	Radar (search & track) guidance	Short-range radar, communications, & guidance
Tower	Whips	Stubs	Stubs	Lens
Whips	Tuned stubs	Biconical	Helices	Slotted helix
Loops	Trailing wires	Partial sleeves	Flat biconical	Horns
Half-wave	Ferrite loops	Cap-loaded stubs	Cap-loaded stubs	Slotted guides
Grounded vertical wire	Scimitars	Helices	Horns	Printed arrays
Trailing wires	Valentines	Rods, plastic and ferrite	Traveling wave	Parabolic dishes
Single wire	Yagi	Turnstile (TV)	Inflatable	Geodesic
				Luneberg lens
Rombic	Spirals	Spirals	Slots	
	Log periodic	Scimitars	Cones	
Luneberg lens	Inflatable	Valentines	Parabolic dishes	
	Extendible	Loaded slots	Luneberg lens	
Log-periodic	Luneberg lens	Corner reflector	Printed arrays	
Discone		Yagi	Slotted wave-guides	
Sleeve				
		Log periodic	Spirals	
Broadside arrays	Loops	Turnstile	Scimitars	
		Inflatable	Valentines	
Endfire arrays		Extendible	Log periodic	
		Wave-guide horns	Turnstile	
		Parabolic dishes		
		Loops		

3) Gain (above isotropic radiator) and front-to-back ratio
4) Beamwidth (azimuth and elevation)
5) Side-lobe levels
6) Polarization
7) Antenna height
8) Aperture size and type
9) Aperture illumination in both amplitude and phase
10) Frequency bandwidth
11) Patterns at harmonic frequencies
12) Physical requirements

3-46. Control Techniques

a. General. The control of interference by antenna systems is primarily an antenna design problem. The radiation pattern and frequency response is a combined function of the antenna, antenna feed lines, and the counterpoise. Each of these antenna-system components should be designed to re-enforce one another in the generation of the desired radiation pattern and frequency response. The design of simple antenna systems is discussed in the Department of the Army Technical Manual TM-11-666, Antennas and Radio Propagation, and is therefore not discussed in this "Guide".

b. Antenna Patterns. Antenna patterns, both receiving and transmitting, may be employed to great advantage in reducing mutual interference. Directional antennas, that have maximum gain in only the direction for which they are designed, can be used to reject unwanted signals. Side and back lobes can be minimized, rejected, or cancelled completely, so that the antenna provides high values of attenuation in all but the desired direction. Antenna gain at harmonic or spurious frequencies can be minimized so that only the desired frequencies are propagated. Multi-array antenna systems may be utilized, instead of single antennas, for improved pattern and frequency control.

(1) Directional antennas. A directional antenna will have maximum gain in the direction for which it is designed, while omnidirectional antennas have uniform gain in a particular

plane. An antenna should form a pattern only in the direction where reception or transmission is desired. Insofar as practicable, directivity of antennas should be utilized to increase signal strength of desired signals and attenuate signal strength of interfering signals. In this way, systems will operate more efficiently and reliably, interference problems will be reduced to minimum, more power can be transmitted into a smaller area, and weaker signals can be received.

(2) Side lobes. Side-lobe reduction can be obtained by improved antenna design and the use of absorbing or reflecting materials around the antenna itself. One method of improving antenna design is to change the feed pattern to minimize spill-over at the edges of the reflector. The reflector, itself, can be shaped to reduce some of the side lobes, and additional reflectors on the antenna used to reduce specific lobes.

(3) Dual-antenna systems.

(a) Auxiliary-channel techniques. The auxiliary-channel technique utilizes a separate, omnidirectional receiving antenna, and associated receiver in conjunction with the primary receiver, to reduce side or back-lobe interference. In order to distinguish between main-lobe signals and side or back-lobe signals, the auxiliary channel utilizes an antenna system with a high side and back-lobe gain characteristic and a low main-lobe gain characteristic as compared to the primary antenna. When an interfering signal arrives from any azimuth direction other than the one in which the main lobe is pointing, the auxiliary-channel response will be stronger than the main channel contribution. Thus, the side or back-lobe interference can be distinguished from true main-beam signals. Once the side or back-lobe signals are identified, they can be eliminated electronically.

(b) <u>Matched antennas</u>. Side or back-lobe interference can be eliminated by the utilization of two matched antennas. The principle of the matched antennas is that the side and back-lobes in the azimuth pattern of the primary antenna can be precisely matched by the lobes of an auxiliary antenna which has no corresponding main beam. If the pattern of the auxiliary antenna overlaps the primary-antenna pattern, the signals in the auxiliary channel can be directly subtracted from those in the primary channel without affecting the main beam. If the phase centers of the primary and auxiliary antennas coincide, then the subtraction may be done in the rf circuits. If the phase centers do not coincide, then the subtraction can be done at the video level to avoid phasing problems. For complex signal densities involving mutiphase signals, multiphase antennas may be employed.

Section VIII. COMPUTERS

3-47. General

Computer circuits may be adversely affected when interference appears in their power, control, or signal leads. Also, computer circuits themselves may act as interference sources. The permissible level of computer-circuit interference is a function of the desired signal levels and the discrimination that the computer offers to interference. The susceptibility of any computer will vary, depending on its design, construction, installation, and proximity to interference sources.

3-48. Digital Computing Equipment

Compatibility is important in a complex system consisting of many electronic units. When a digital computer is placed in such a system, its operation should not interfere with other equipment, and it should be insensitive to the surrounding electromagnetic environment. To achieve this compatibility, a solid-state digital computer should be thoroughly tested to determine generated interference and interference-susceptibility characteristics. The predominant interference generated by the computer can be attributed to the repetitive operation of the circuits. The interference spectrum is directly related to the organization of the machine, its logic functions, timing, and basic circuitry. Included among the major interference sources are:

1) Master clock timing and basic operations
2) Central processing system and programmed operations
3) Core storage elements
4) Drum system and associated processing circuits
5) Typewriter, card reader, and associated processing circuits

3-49. Analog Computing Equipment

Because analog computers operate with conventional wave-shapes, they are susceptible to the effects of interference fields from either repetitive, pulsed, or random noise. Computers used for process control are generally located near power equipment. In such situations, the design engineer should expect such interference as:

1) Transient fields from solenoids and similar components, with high inrush current affecting counter circuits
2) Fields emanating from power transformers and filter reactors
3) Stray fields from magnetic amplifiers in power servos affecting servo accuracy

3-50. Interference Sources

Interference sources, that can be minimized by proper system wiring, may be divided into two groups: those that originate inside the data system and those that are external to the system.

a. Inside the Data System. Internal interference sources that can be minimized are dependent upon various forms of unintentional circuit coupling. The most common form is common-resistance coupling in the ground scheme. Common-resistance coupling occurs when current in one circuit causes a voltage to appear in the other circuit(s), when two or more circuits share any resistance, or when they connect at a common point. The circuits, in any system, exist as closed loops that have mutual inductance. This inductance varies directly with the area enclosed within the loops. When a current change occurs in one of these circuits, it causes voltage, due to the mutual inductance, to appear in the other circuit(s). The significant parameters are the area enclosed by the circuit, its impedance, and the relative power levels. Every portion of a system has capacitance between itself and every other portion. Changing voltages, regardless of location, tend to drive current through these capacitances and produce interference.

b. External to the System. The external environment of a system can cause malfunction. The effects of electrostatic and electromagnetic fields can be controlled by the same methods used to minimize the effects of internally-produced interference. In addition, system wiring can often be protected from external fields by proper enclosures and packaging.

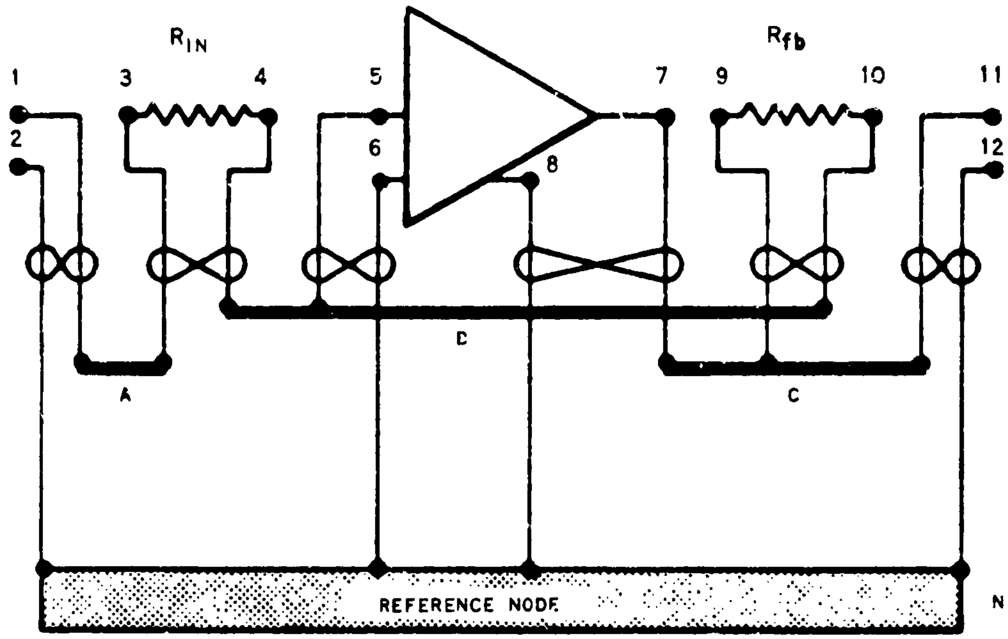
3-51. System Nodes

a. General.

The data system is designed to transfer the potential between one pair of nodes to another pair of nodes without introducing error. Power must be transmitted between the pairs of points to maintain the signal above inherent noise; the current flow must cause a potential drop along the connection. The pairs of nodes, in a system between which voltages are measured, are not all independent. Most such node pairs have one reference node (ground) in common. In a complex system, the reference node is needed at many widely-separated locations. Its configuration is a major problem, aggravated by the widely-varying power levels of the circuits that connect to it. To avoid mutual impedance coupling in the leads going to the reference node, the node should be a single point. All ground connections in the system should be made at this point. As a practical consideration, some resistance can be tolerated in the reference node, depending on the maximum interference current, and the allowable input interference voltage of the most sensitive circuit. A finite volume can be used for the node. For practical minimum resistance, the material should be solid copper. All system voltages are understood to be with respect to the reference node. Because it is constructed to allow no significant voltage drops, the node structure provides a true equipotential reference. Although a number of connections may be made to this reference node, the requirements of negligible mutual resistance necessitate that the node structure be as compact as physically possible. These requirements generally produce a node structure in the form of a closely-packed rectangle. In an extremely large system, the node structure may take the form of the five sides of a cube. In any case, the reference-node structure must be large enough to permit all ground connections in the system to be made on the block.

b. Construction.

(1) Each system node presents the same problem as the reference node in greater or lesser degree, depending on the number of branches. Mutual-impedance coupling can occur in any node if it is not a true ground point; however, no system node has nearly as many branches as the reference node. The system nodes are usually of similar construction to the reference node, but much smaller. The wires, connecting the modules to various nodes, should be grouped into cables, each containing a complete circuit. When the conductors in each cable are twisted, the area enclosed by the circuit will be reduced. Such a cable will not be subject to interference from external magnetic fields; conversely, such a cable will not generate substantial external magnetic fields. Thus, minimal electromagnetic pickup and minimal noise-producing fields will result. Figure 3-159 illustrates the application of this method. The reference node is indicated by N. Adjacent to it are the various system nodes (A, B, C). The amplifier is shown with resistors R_{in} and R_{fb}. The amplifier input ground (point 6), and the signal ground (point 2), are connected directly to the reference node. If the inputs were wired point to point (point 1 to point 3, then point 4 to point 5) instead of as shown, the area enclosed by the circuit (1, 2, 5, 6) would be much greater than if the system nodes are used as indicated. Any attempt at point-to-point wiring between portions of the system will result in large area loops being included in many circuits. Because the susceptibility of a circuit is directly proportional to the area enclosed in the circuit, point-to-point wiring should not be used in conjunction with a reference-node structure.



IN1212-63

Figure 3-159. System Nodes and Reference Node - Operational Amplifier

(2) All system nodes should be located very close to the reference node. In general, this means that none of the system nodes should be at the normal terminals of the component portions of the system. The terminals of these various component portions should be connected to the system nodes and reference node through lengths of wire. Because the only nodes are at the system reference-node structure, the wires are considered as part of the internal component construction. The terminals of R_{in} (points 3 and 4 on figure 3-159) are of no significance in the system. They must be wired to system nodes A and B, and the resistance of the wire considered part of R_{in}. Similarly, the terminals of R_{fb} are wired to nodes B and C, and the resistance of the wire becomes part of R_{fb}. The wires are then twisted together, as indicated, to form cables containing each complete closed circuit. Feedback is taken directly from the

system node structure. The wire resistance is included inside the feedback loop, and the amplifier output impedance is not substantially increased by the use of long wires leading to the node structure. The amplifier output exists at the node structure, and the voltage at the actual amplifier terminals is of no more significance than any other internal amplifier voltage.

(3) Potentiometric connection of the same amplifier is shown on figure 3-160. Voltages at the actual amplifier terminals (points 5 through 8) are not meaningful. The potentiometric amplifier input nodes are A and B, and the amplifier output nodes are node C and the reference node. If wiring is done as shown on figures 3-159 and 3-160, it is of the utmost importance that the amplifier input low side (point 6) and the amplifier output low side (point 8) not be connected internally at the amplifier. If such an internal connection should exist, connecting both these points to the reference node would create a ground loop. Such a loop would be a complete circuit and may enclose a substantial area. This loop would be subject to pickup from any changing current in the system, or any magnetic field existing in the region where the system is located. Unfortunately, points 6 and 8 are often connected internally. When this is done, only one point should be connected to the reference node. Single wires, from the amplifier-input high side and the amplifier output, should be run together with the single ground wire as one cable. All area loops, whether due to multiple connections to the reference node or to any other node, should be avoided.

(4) The twisted cables, connecting the various parts of the system to the node structure, are seen on figures 3-159 and 3-160. The building of the node structure eliminates all common-resistance coupling between circuits of the system; and the use of twisted cables reduces the susceptibility of these circuits to electromagnetic pickup.

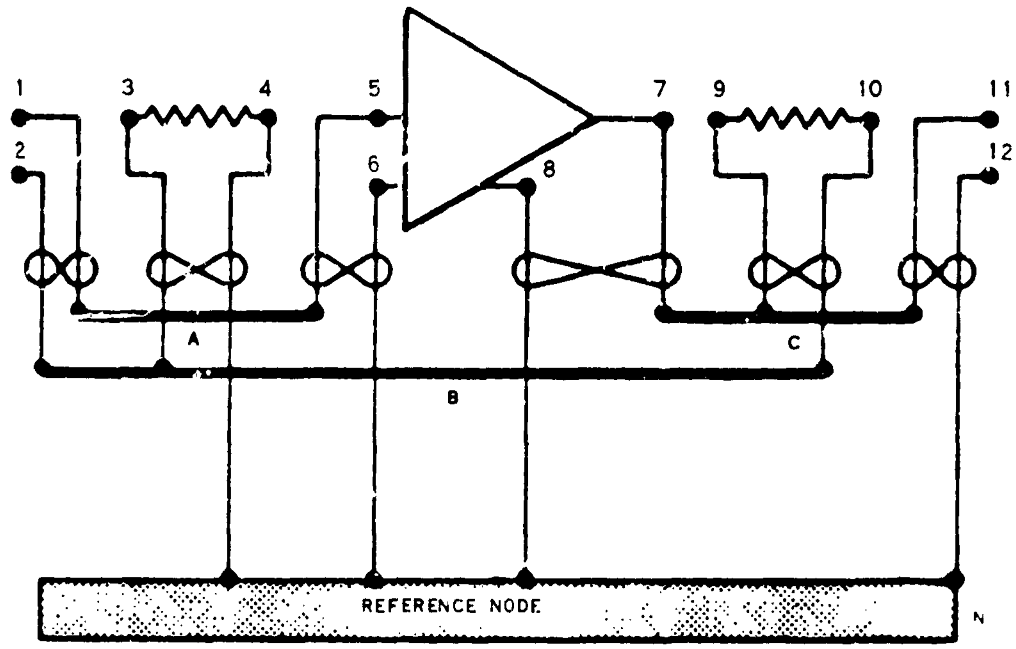


Figure 3-160. System Nodes and Reference Node - Potentiometric Amplifier

c. Cable Considerations.

(1) The grouping of system cables around the reference node provides an easy point of connection for any required shielding. Should it be desirable to shield any cable that has at least one wire tied to the reference node, the shield can be connected to the reference node. Any cable, that does not have a wire to the reference node, should have its shield connected to one of the system nodes at which it terminates -- ordinarily, a node that has the least impedance to the reference node. All shields should be connected to a node that is at the same potential as the circuit that they are designed to protect. The following rules apply:

1) Establish a reference node structure

2) Establish a complex of system nodes around, or upon, the reference system node structure

3) Make system interconnections only at the reference-system node structure (do not use point-to-point wiring and do not ground the other end of a wire that ties to the reference node)

4) Connect a conductor to any node only once

5) Group all connecting wires into twisted cables, each of which contains a complete unit

6) Enclose, in shields, any cables that carry power or rapidly changing voltages of any type, or that are susceptible to electrostatic pickups

Allowable exceptions to these rules are:

1) Circuits, not susceptible to noise, may be wired without connection to the reference-node structure; however, the rules regarding twisting of complete circuits and shields should be followed; otherwise, these circuits will establish changing fields in other circuits

2) In many types of high-speed circuits, the capacitance, involved in the long twisted cables called for by this system, is not tolerable. In these circuits, interference rejection may not be supplied by twisting the leads

(2) For greatest protection against outside interference, the supporting structure of the system must ordinarily be connected to the system reference node. In addition, the chassis of the various modules are often tied to system ground. Unless all modules in the system are properly isolated from the supporting structure, ground loops may result. In some cases, capacitive coupling to the supporting structure may cause ground loops through which high-frequency currents can circulate. These currents may necessitate surrounding various system components with shields. The result of following the rules is a computer that

appears to have a large number of excessively long interconnections. The long cables, however, are merely extensions of the various component parts of the system. No system nodes exist at the terminals of the actual components. Interconnections are made only between the system nodes, which are closely grouped at the system central ground location. The interconnections in this system are actually only those very short lengths of wire between the system nodes.

d. Components.

The wiring recommendations made here place severe limitations on the computer components that may be used. Among the most important of these components are power supplies and amplifiers.

(1) Power supplies. Power supplies, for use in susceptible portions of a computer that employs a wiring system based on recommendations made here, must be suitable for use with the reference-node structure. The regulated power supplies must be remote referencing. To hold the system-node voltage constant with respect to the reference node, the feedback must be from the system node. Unregulated power supplies should have their final filter capacitor connected directly across the system nodes to avoid an increase in output impedance from the external wires. Multiple-voltage power supplies should not have a common internal ground connection. To prevent the unintentional creation of ground loops, power supplies should be isolated from the power line by carefully-shielded power transformers.

(2) Amplifiers. Feedback amplifiers, for use in a system using the wiring methods outlined, must be capable of remote feedback connections, as indicated on figures 3-159 and 3-160. If the feedback is taken from the module terminals, the resistance of the wires connecting the amplifier to the system nodes becomes part of the amplifier output impedance. A further problem, in

connecting the amplifiers in the system, is the difficulty of determining the proper number of wires to use. If the amplifier is a true four-terminal device with an external, isolated power supply, three cables will connect to it: input, output, and power. If the amplifier is really a three-terminal device, this type of connection cannot be used. Further complications may be introduced when the power-supply ground is common to either the input or output low side of the amplifier, or both. Figure 3-161 shows proper connections to amplifiers of these various internal configurations.

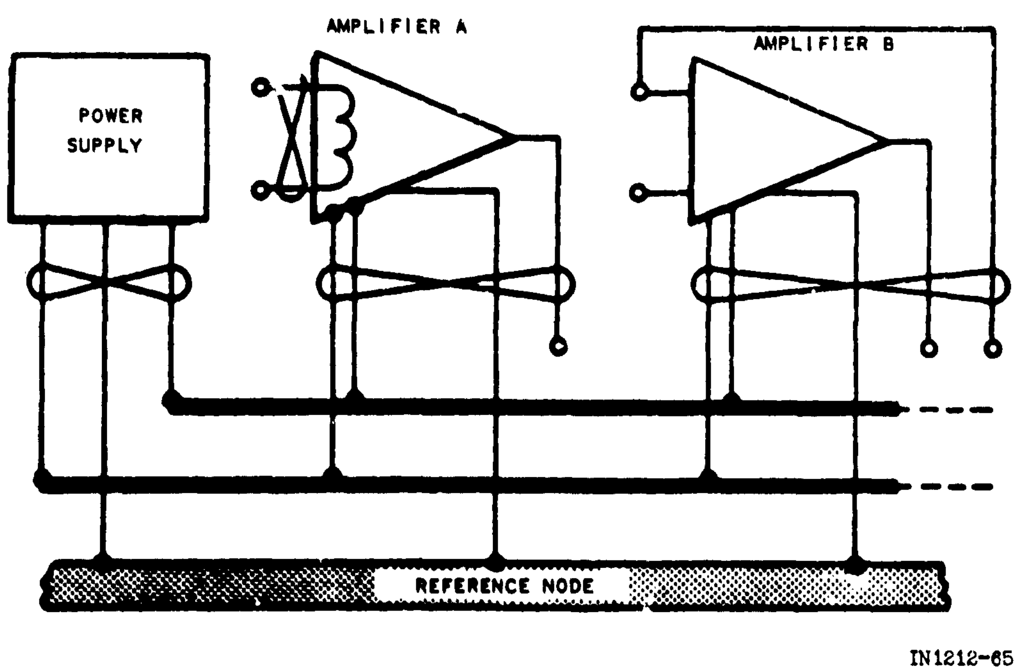


Figure 3-161. Power Supply Connections for Three-Terminal and Four-Terminal Amplifiers

Section IX. MICROMINIATURIZATION

3-52. Introduction

The objectives of electronic microminiaturization for equipment are increased reliability, and reduction of size, weight, and cost. Size, weight, and cost reductions are achieved by new fabrication techniques, such as electrolytic deposition or evaporation of thin films. Increased reliability is the result of fewer solder connections, replacement of solder connections by chemically-bonded material interfaces, improved control of materials and processes used in fabricating circuit elements, and increased redundancy at component or circuit levels. Miniaturized equipment is less susceptible to external interference than conventional equipment because of the smaller areas and volumes available for interference coupling, and the ease with which smaller volumes can be shielded. Also, because of the low-level operation of miniaturized equipment, the generation of radiated and conducted interference is minimized. There are two interference problems in microminiaturized packaging: interference between microcircuits, and interference that originates external to microcircuits. Microminiaturized equipment operates at low-power levels, so that the radiated energy is negligible and does not cause interference. Problems exist when only a part of a system is miniaturized. In such a case, the electronic design engineer, responsible for interference reduction, should decide if the remaining nonminiaturized portion will affect the operation of the miniaturized unit. The engineer should become familiar with the different types of resistors and capacitors available for miniaturized circuit design.

3-53. Basic Miniaturized Components

a. There are three typical resistors used in microminiaturized circuit design; two of them are shown on figure 3-162. The first type is simply a bar of semiconductor with ohmic contacts at both ends (fig. 3-162A). This bulk type is limited to low values of resistance. One reason for this low resistance is that the bar can not be too thin or narrow, or its lack of physical strength will make it difficult to handle. The second resistor type, the diffused-layer silicon resistor, is shown in different versions on figure 3-162B and 3-162C. A thin layer of p-type material is formed in an n-type bar. When reverse biased, the p-n junction isolates the diffused layer from the rest of the bar. Because the diffused region can be very narrow and thin, the resistance of this configuration is greater than that of the bulk resistor. The third resistor type is the thin-film deposited resistor which is a resistive material deposited on an insulating substrate. The resistive material can be chosen from a wide range of materials, varying from common nichrome to exotic, specially-fabricated alloys. Large values of resistance are possible, using the thin-film technique, because the deposited layer can be made small in cross-sectional area.

b. There are two basic types of capacitors for use in miniaturized circuitry: the thin-film type and the p-n junction type. The thin-film capacitor is composed of two conducting layers with a film of dielectric between them. Common configurations are successive layers of silicon, silicon oxide, and a deposited metal, or successive layers of deposited metal, an insulator, and then another layer of deposited metal. The p-n junction normally operates as a diode; but, if the junction is reverse-biased, it functions as a capacitive element.

c. The resistive and capacitive circuit elements can be combined to form distributed RC networks, as shown on figure 3-163. The diffused layer shown has high conductivity; the p-n junction supplies the capacitance, and the bulk material functions as the resistance. Various electronic functions can be obtained from the same circuit by different contact arrangements. For example, it is possible to have one contact on the bottom instead of two or three contacts.

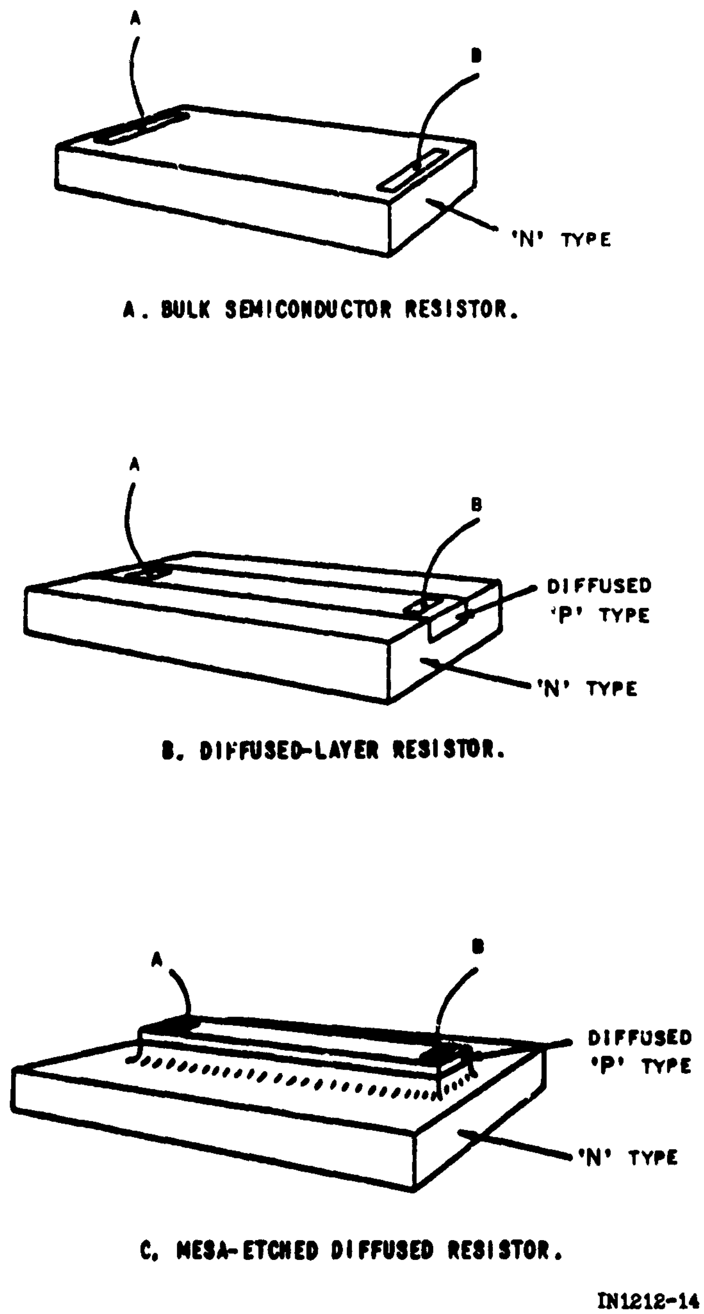
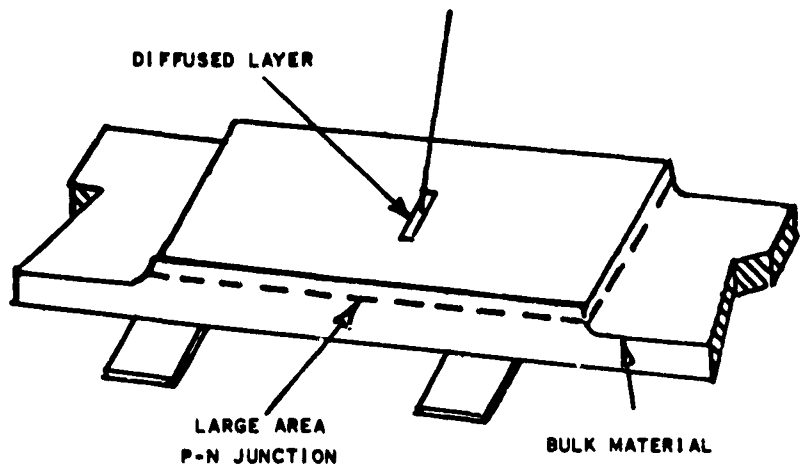
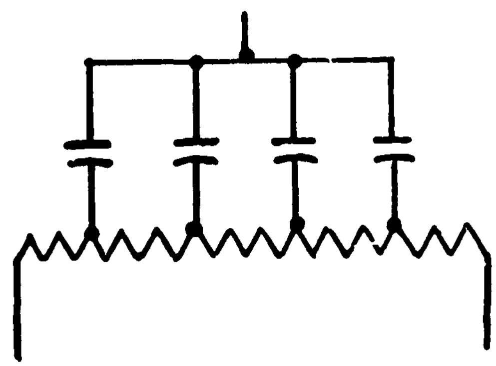


Figure 3-162. Semiconductor Resistors for Microminiaturized Circuit Design



A. ASSEMBLY OF DISTRIBUTED RC NETWORK.



B. SCHEMATIC OF SEMICONDUCTOR DISTRIBUTED RC NETWORK.

IN1212-15

Figure 3-163. Distributed RC Network with Equivalent Circuit

d. There is no satisfactory way to produce miniature inductors. A circuit to be miniaturized should be completely redesigned with a view toward eliminating all of the inductances. Although this is not always possible, a significant reduction in the number of inductors used in typical circuits can be effected.

3-54. Miniaturized Circuit Formulation

There are three basic approaches used to form miniature circuits: discrete components, multilayer thin-film, and semiconductor solid-state.

a. Discrete-Component Circuit. In this approach, each component is produced separately and miniaturized individually. In the micromodule system, each component is placed on an individual ceramic wafer, and then several wafers are stacked to form a module. The standard micromodule dimensions are 0.31 x 0.31 inch, with the wafer thickness allowed to vary depending upon the component. On each wafer, connections are brought out to notches on the sides of the wafer. Riser wires are placed in the notches to interconnect the components and form the desired circuit. After the connections are soldered, the entire unit is encapsulated in epoxy for protection. The finished module may be plugged into a printed-board circuit in the same manner as a transistor. The height of a module depends upon the number and type of components used. Another discrete component design involves placing a complete circuit inside a transistor can. Because of the resulting small volume, it is unlikely that an interference signal could be induced in one of the internal circuit-loops. The largest loop in this type may have an average area of only 0.002 square inch, which precludes interference coupling. These circuits,however, are interconnected by printed circuit-boards, or similar devices, and therefore produce larger loops and have increased interference susceptibility.

b. Multilayer Thin-film Circuit. Components, such as resistors and capacitors, can be readily fabricated by thin-film techniques, both accurately and reliably. Inductors, transistors, and diodes have not as yet been successfully produced by this method. Numerous thin-film resistors and capacitors can be deposited on top of one another to form complex-circuit multilayer structures. Insulating layers,deposited between the components,provide isolation.

Figure 3-164 represents 16 layers, formed coincident with each other, to produce a typical digital circuit. All the necessary interconnections are also supplied by the depositions. The complete structure is only a few mils thick. Multilayer structures can be made small enough to be effectively used as a single component in discrete component circuits. In these multilayer structures, magnetic coupling is not much of a problem, but the capacitive coupling effect can be pronounced. Coupling within a single structure is significant and, therefore, should be taken into consideration in the initial design of the structure. Coupling between adjacent structures can be substantial; therefore, care must be used in the physical arrangement of the system.

c. Semiconductor Solid-State Circuit.

(1) A semiconductor solid-state circuit requires that all circuit components be located in, or on, one block of silicon. The desired components are formed and interconnected by a series of diffusions and deposits. This type of construction has the advantages of smallness and reliability. The increase in reliability is due, in part, to fewer and simpler interconnections. These solid state circuits have two serious drawbacks: the circuit elements have loose tolerances because of fabrication limitations, and the production line yield is low (with a corresponding high cost).

(2) A typical NOR semiconductor solid-state circuit is shown on figure 3-165. The schematic circuit is shown in figure 3-165A; the B circuit is similar to A, but is shown in three dimensions to simplify the explanation of C. The individual resistor, R_A, and capacitor, C, in schematic A, have been changed to a distributed network in B. The actual semiconductor solid-state circuit is represented by figure 3-165C. In C, large bars are composed of solid silicon and used as bulk resistors. The four round dots

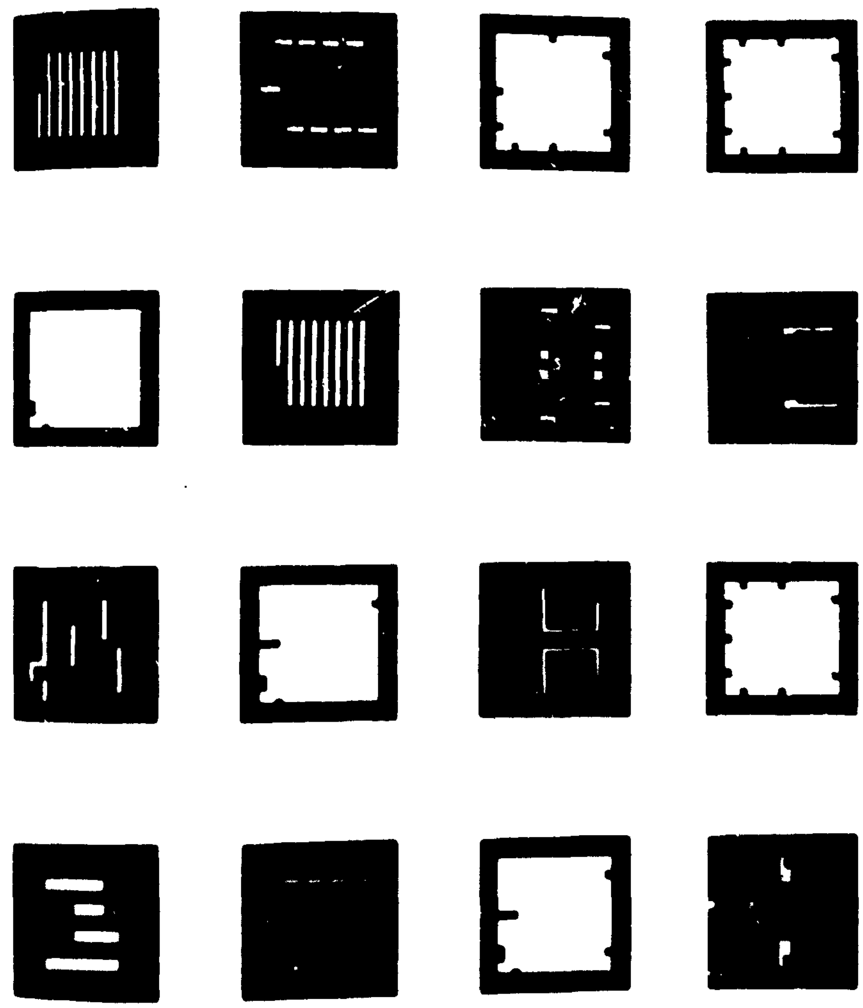


Figure 3-164. Sixteen Evaporated Layers Forming a Typical Digital Circuit

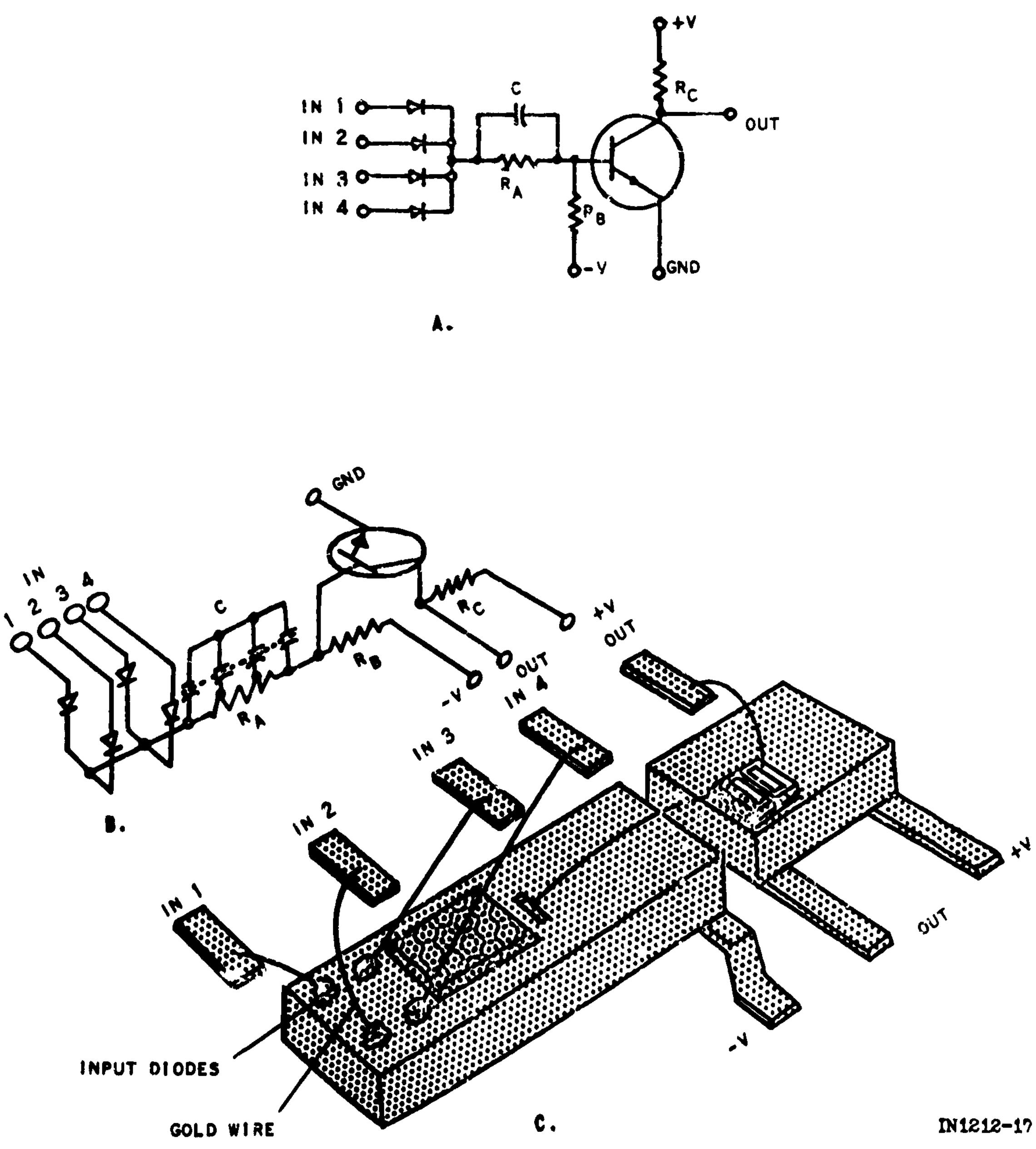


Figure 3-165. Typical Semiconductor Solid-State NOR Circuit

to the left are the input diodes, which are formed on the bar by diffusion. A fine gold wire connects each diode, to the appropriate input lead. To the right of the diodes is the distributed RC network. The large, rectangular, raised area is a silicon-oxide layer with a metal electrode on its surface so that the result approximates the distributed network shown in B. To the right of the distributed network is a small rectangular ohmic contact, which is a small area of conductor alloyed to the silicon bar to provide a lead bond at that point. A gold wire runs from this contact to the base of the transistor, as shown in B. The resistor, R_B, is the bulk resistance between the ohmic contact and the tab at the bottom labeled (-V). The transistor is formed and located on the shorter bar. The remaining component, R_C, is the bulk resistance between the collector of the transistor and the tab labeled (+V).

(3) Another example of a semiconductor solid-state circuit is illustrated on figure 3-166. The schematic circuit diagram is A; C is the same circuit redrawn with the individual resistor and capacitor replaced by a distributed network. The model at B shows only the bulk resistors; the completed circuit is illustrated at D, which shows two transistors, gold wires, and the deposited layers that form the distributed RC network. This circuit is enclosed in a package, 0.250 x 0.125 inch in outside dimensions. For interference susceptibility, the most critical loop contains the emitter and base of a transistor. In this circuit, the area of the loop is estimated to be 0.0003 square inch; therefore, the possibility of interference is remote. Larger, more critical loops are formed when such circuits are interconnected. This is not serious considering that, because of the over-all size reduction,

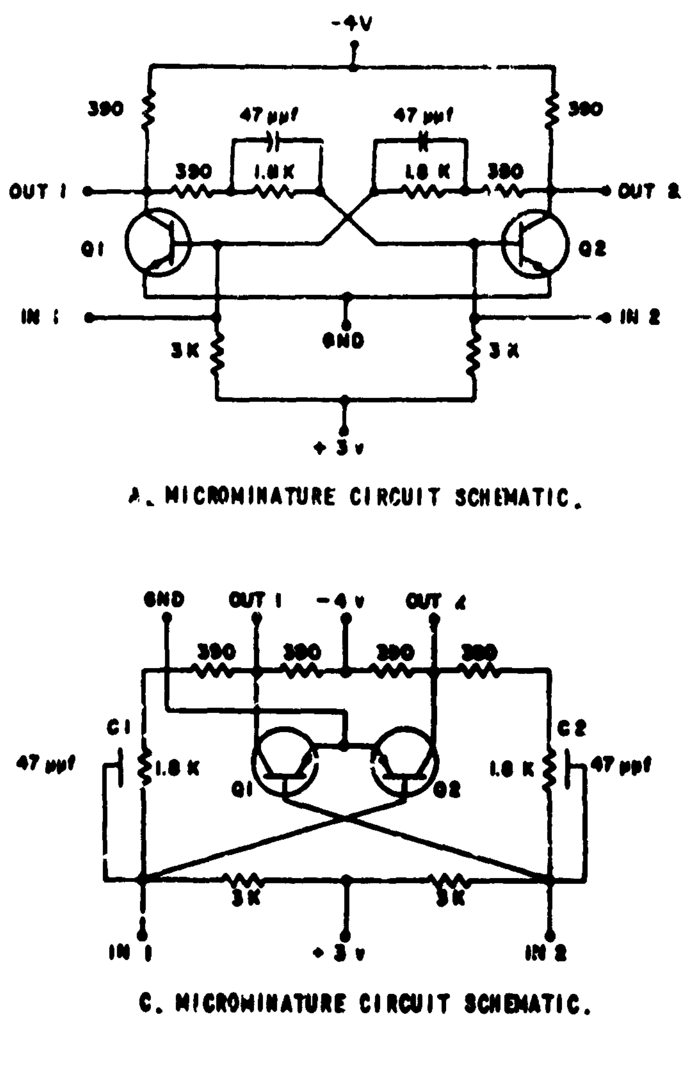


Figure 3-166. Typical Semiconductor Solid-State Flip-Flop

the loop in a solid-state circuit will only be about 1 percent of the area of a similar loop in conventional equipment. Miniaturized components, in mechanical proximity to each other and subject to magnetic coupling, can be wrapped with strips of Netic or Co-Netic foil, or its equivalent, prior to final assembly. Such wrapping eliminates special tooling of small enclosures to accommodate each of the affected items. In many applications where miniature components are used, a wide variety of enclosures can be fabricated to protect the components.

3-55. Miniaturized Circuitry

Miniaturized circuitry has been widely used in if amplifiers. The if strip module, shown in figure 3-167, has several fabrication characteristics typical of miniaturized circuits. These include:

1) Resistance-welded joints between component leads and longitudinal interconnecting nickel ribbons

2) Polyurethane foam potting-compound to support both joints and components

3) Nickel-sheet shield-can, 0.005-inch thick, to provide a tolerance-free potting mold and interference shielding for the circuitry

4) Coaxial-type sockets flush with the module shield-can, or nickel-ribbon leads, or pins, extending from the module shield-can for interconnection

5) Shield-can cover, soldered on with nickel peel-strip for ease in removing cover

6) Tuning holes provided in the shield-can. Although not really a part of the module design, a soldered tab system is essential for mounting. This system eliminates hardware and simplifies installation or removal

a. Figure 3-168A shows the schematic of a 24-mc if strip with four stages, 80-db gain, and approximately 2-mc bandwidth. The actual construction used is shown on figure 3-168B. Interference can be reduced if all the leads are shortened to minimize their acting as radiating sources and receivers. The stages are transformer-coupled, therefore, provision for tuning after final assembly must be provided. This type of circuitry is also susceptible to short-term, electromagnetic pulse radiation. The nickel shield-can is the answer to both the interference and magnetic problems. The stages are separated by nickel shields to provide some feedback reduction.

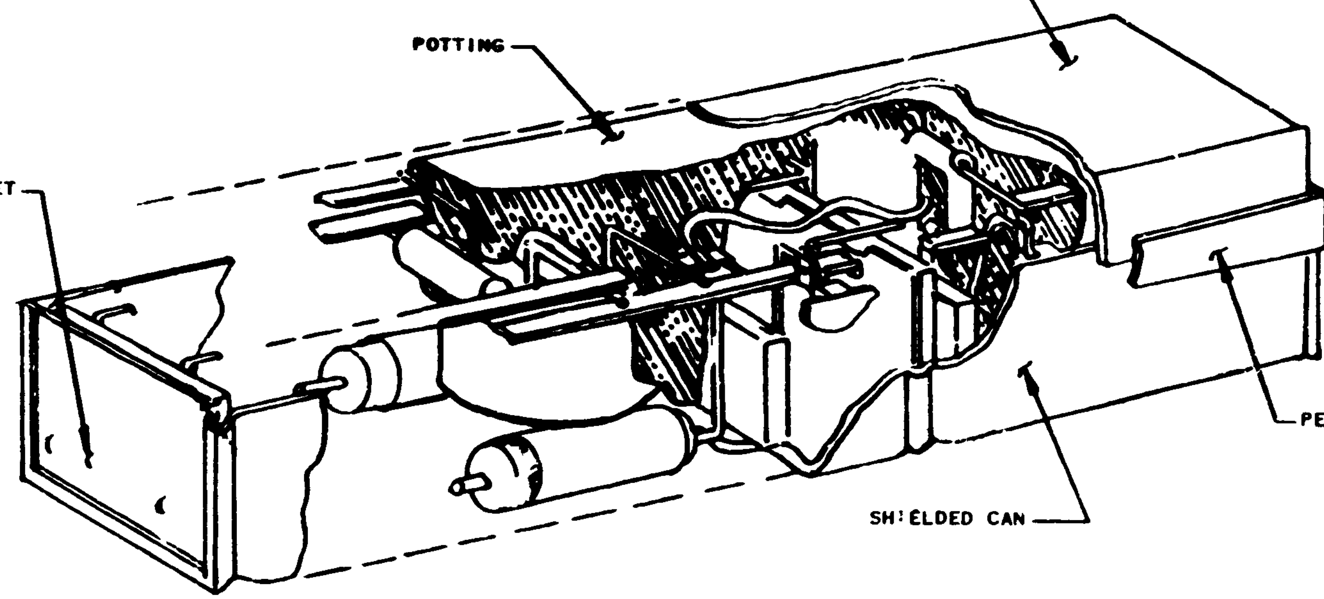
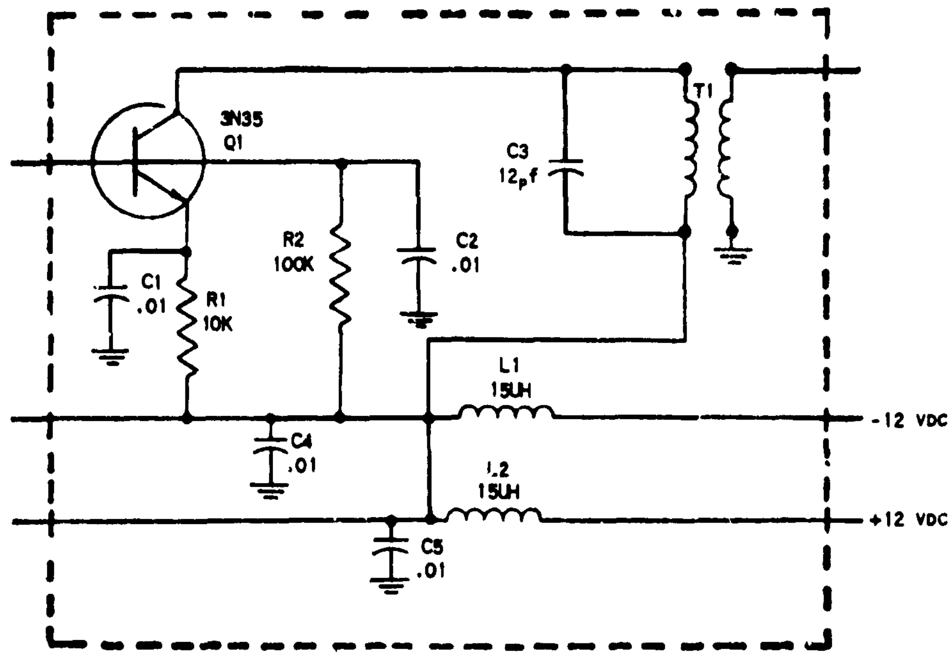
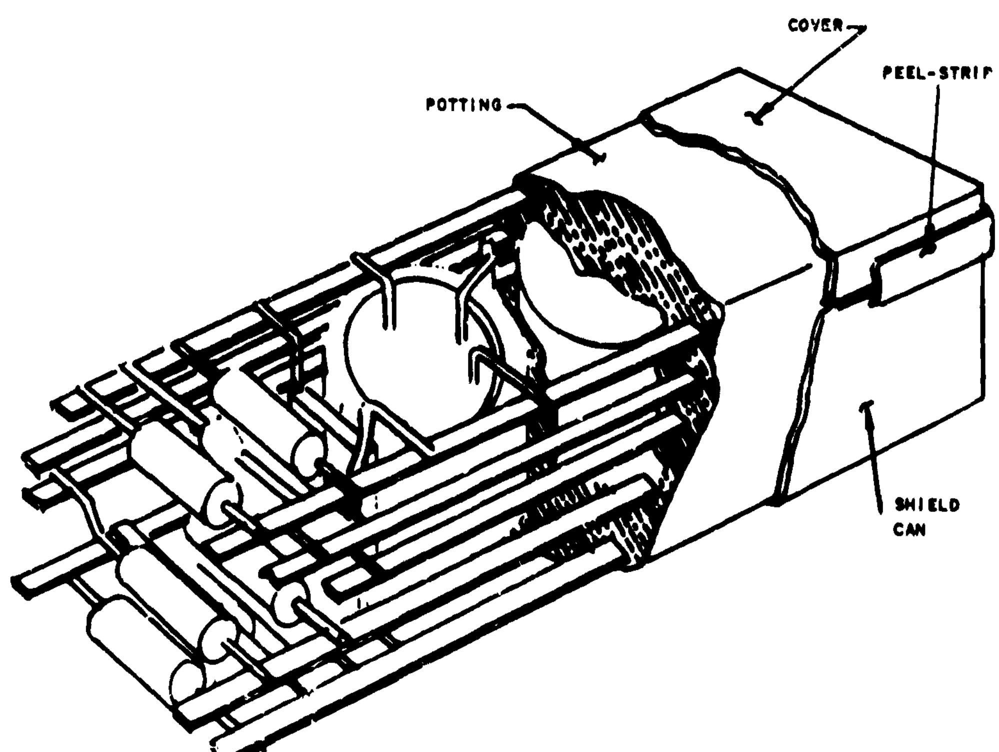


Figure 3-167. Typical IF Strip Module Construction



A. SCHEMATIC OF I F STRIP



B. CONSTRUCTION OF I F STRIP

Figure 3-168. Typical Four-Stage IF Module Construction

b. The module contains 60 standard components. They are arranged in four levels to simplify assembly. The four levels of a single stage are placed on individual welding jigs. The components for three stages are placed in the jigs before further steps are taken. The figure shows nickel-ribbon interconnecting wire for the circuit. These ribbons run the length of the module as common or bus connections, or can be cut at appropriate points to perform separate circuit functions. The four levels are assembled and welded, and intermediate ribbons are cut and inspected separately. The four levels are assembled first into two groups. Interlevel welds are made, and the two major groups assembled and interwelded.

c. In some design techniques, shield planes are designed within the wiring module (fig. 3-169). Shield planes are also used to distribute dc potentials to the individual subassembly modules by means of the through-wiring technique. The -4 volt dc and -10 volt dc planes act as shield planes since they are effectively connected to ac ground by the 3.3-μf tantalum capacitors. These connections are satisfactory, provided that the inductance of the lead lengths of the capacitor, the inductance of the capacitor, and the capacitance, are of such value, at the frequencies of interest, to provide a low-impedance ac connection to ground.

3-56. Ceramic-Filter Application

a. Ceramic filters enable the electronic design engineer to depart from conventional design-limiting considerations. The increased demand for small electronic components has made the miniature ceramic ladder-filter very desirable. No magnetic shielding is required with these filters.

b. Receiver design is simplified when ceramic filters are used. Many space-consuming if stages can be eliminated and improved selectivity still be obtained. In single-sideband communications, ceramic ladder-filters are very effective in rejecting the carrier frequency and the undesired sideband. Ceramic bandpass filters often eliminate an extra stage of conversion.

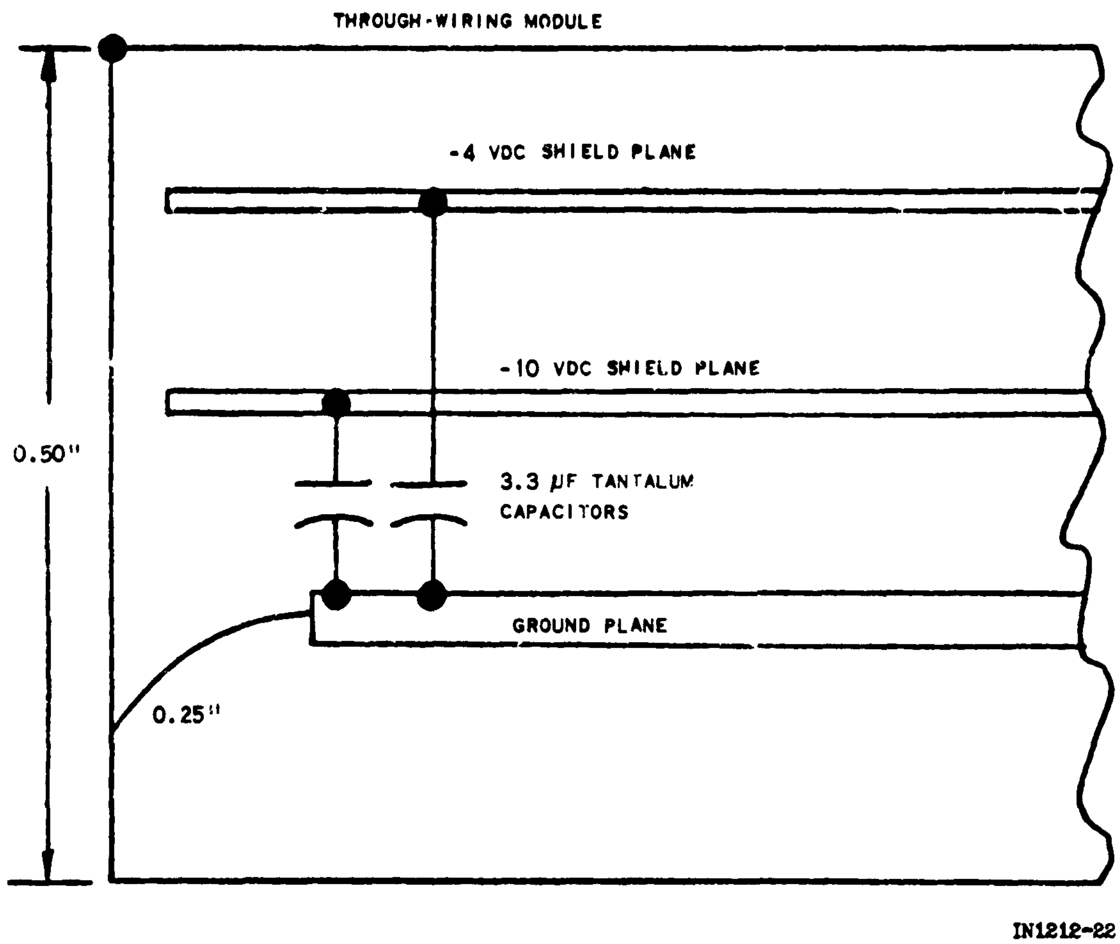


Figure 3-169. Shield Planes for Distribution of DC Potentials to Individual Subassembly Modules

c. Ceramic filters are ideally suited to transistor applications where filter impedance must be in the order of 1000 to 2000 ohms. Since the effective Q of piezoelectric materials is higher than that obtained from electrical components, it is possible to transmit energy through ceramic filters, over a band of frequencies, with a lower insertion loss. Using materials having Q's ranging from 50 to 2,000, bandwidths are obtainable that range from 1 to 10 per-cent of center-frequency. The power insertion loss of these designs, which is dependent on both Q and bandwidth, ranges from 0.5 to 15 db. Insertion loss of ceramic filters is lower, and their skirt selectivity is better, than that of electrical filters. Ceramic disc-type ladder filter designs suppress spurious responses 60 to 100 db below the passband level. They may be employed, for most band-pass filtering requirements, in the frequency range from 100 to 1000 kc, and are especially applicable for use in carrier systems and single-sideband equipment, as well as communications receivers with high-performance requirements. Ceramic filters are well suited to transistor circuit applications because of physical size and low impedance.

d. In radio receiver applications, a piezoelectric filter fulfills the need for selectivity in a fashion superior to that usually provided by a multiplicity of if transformers. Not only does the ceramic filter provide a flat passband characteristic and steep skirts on the selectivity curve, but, in addition, its availability as a lumped selective network allows the design engineer considerably more freedom in providing an optimum receiver design. The stability of ceramic filters makes them highly desirable for use in communications receivers. Because of aging, receivers using electrical filters experience filter-characteristic changes after one year's service. Such aging requires that the receiver be realigned and a new frequency crystal placed in the circuit. Ceramic filters can operate for ten years before any such realignment is necessary. In the 400 to 500 kc frequency range, development has now reached the stage where temperature and aging properties of the ceramic materials are the limiting factors in their use. Work on the temperature stability of materials has progressed to the point where temperature extremes of 150°C can be tolerated. For higher operating temperatures (200° to 250°C), further work will be necessary to improve the stability of ceramics.

e. Manufacturing processes used in ceramic filter fabrication result in low-cost production; they are less expensive to produce than electrical filters. With large scale production and continuing improvements in techniques, the cost of ceramic-tuned circuits should decrease appreciably.

Section X. ELECTRICAL MACHINERY

3-57. General

Interference reduction design for electrical machinery is divided into four categories: interference reduction for large motors and generators, for alternators and synchronous motors, for fractional-horsepower machines, and for special-purpose rotating machines. The electronic design engineer should regard any rotating machine with sliding contacts as a potential source of interference because the switching and arcing processes of commutation cause rapid current and voltage changes that distribute interference energy throughout a wide frequency range.

3-58. Brushes

Brushes and brush leads are the most likely components from which interference can be radiated or transferred. If a motor or generator is not completely enclosed in such a way as to be adequately shielded, then the brushes and brush leads may require shielding. The electronic design engineer should insist that provision be made in the original design of motors or generators for installation of capacitors at the brushes. Brush-generated interference may be reduced by incorporating the following in the design:

a. Brush Pressure. Generated interference decreases at all frequencies with increasing brush pressure. Increased brush pressure, however, increases the rate of wear. The necessity, nevertheless, of more frequent brush replacement is a reasonable compromise for the sake of decreased interference.

b. Current Density. Generated interference decreases with decreased current density. As the current density is increased, more heat is generated at the brush surface,sliding on the commutator or slip ring. This heat causes the formation of a thick oxide film on the sliding metal surface. Rapid variations in the sliding contact resistance,resulting from irregularities in this oxide film, cause high-frequency transients that produce interference. To offset the heat increase, a somewhat larger brush-surface area than necessary should be designed. Such a design change will reduce heat and losses due to mechanical friction. On the other hand, if too low a current density is used, nonuniform grooves develop on the metal surface of the slip ring or commutator, and frequently the increased friction, due to the wider brush-surface area, sets the brushes into a noisy chatter.

c. Brush Resistivity. Brush materials of low resistivity are poor interference generators and are therefore desirable for use in interference reduction designs. A good example of such a brush is an electrographitic carbon brush with 0.0015 to 0.0025 ohm specific resistance in machines being used at less than 50 volts. Low-resistance brushes are available with silver, copper, or cadmium impregnated graphite. When used with a commutator, the resistance of the brush should match the requirements for good commutation. The design engineer should select material of the lowest resistivity that satisfies the other requirements of good functional performance. When used with slip rings, a wide choice of brush material is available because no switching action is involved.

3-59. DC Motors and Generators

a. Of all rotating machinery, dc motors and generators are the most serious offenders in generating interference because they require commutators for their operation. Commutation is essentially a switching action that is accompanied by interference-producing transients and brush interference. When a switch is closed in an electrical circuit, the input impedance changes from practically infinity to zero. If the circuit contains inductance and/or capacitance, its voltages and currents cannot return to normal instantaneously because the energy, stored in the magnetic field of the inductance (or in the electric field of the capacitance), cannot change instantaneously. Initially, the changing voltages and currents develop steep wave-fronts, rich in harmonics, which decay as a function of time. The bars of a commutator, sliding rapidly past the contacting brushes, produce a switching action. This action causes extreme variations in impedance, which, in turn, establish the series of voltage transients, or pulses, that cause interference. Measures can be taken, in designing a generator, to minimize the amount of interference generated by commutator action. Reduction of commutation transients requires the use of design techniques to provide a smooth transition from one value of impedance to another within each armature coil. Interference, produced as a result of commutation, is reduced by five design techniques: interpoles, compensating windings, increased number of armature coils and commutator bars, laminated brushes, and commutator plating.

b. The best way to improve commutation is by adding interpole windings. Interpoles counterbalance the self-induction of the armature coils during the commutation period, and also reduce the induced voltage in the armature coils resulting from the coil-cutting fringing-flux during the commutation period. The use of properly designed interpoles produces a rapid change in the armature-coil current at the beginning of the commutating period, reducing the steepness of the transient at the end of the commutating period.

c. Compensating windings produce, to a lesser degree, the same effect as interpoles and, in addition, help to prevent field distortion. They also assist in reducing cross flux produced by the armature coils. The use of interpoles and compensating windings lessens critical brush positioning requirements with respect to the commutator, and provides electromotive forces in the coils under commutation which oppose the electromotive forces of self and mutual induction in these coils. Increasing the number of coils on the armature (thereby increasing the number of commutator segments, or bars) reduces interference by reducing the current broken per bar and the reactance voltage per coil. The largest number of armature slots, in which the coils are uniformly distributed with respect to the commutator bars, should be used, and the armature slots should be as shallow as possible. The use of short-pitch windings reduces interference by reducing the reactance voltage of each coil. The break transients, resulting from the switching action of the commutator, can be smoothed out through the use of laminated brushes. These consist of brush materials of different resistivity, cemented together by nonconducting glue which provides insulation between adjacent brush segments. The ideal operation of laminated brushes is indicated on figure 3-170.

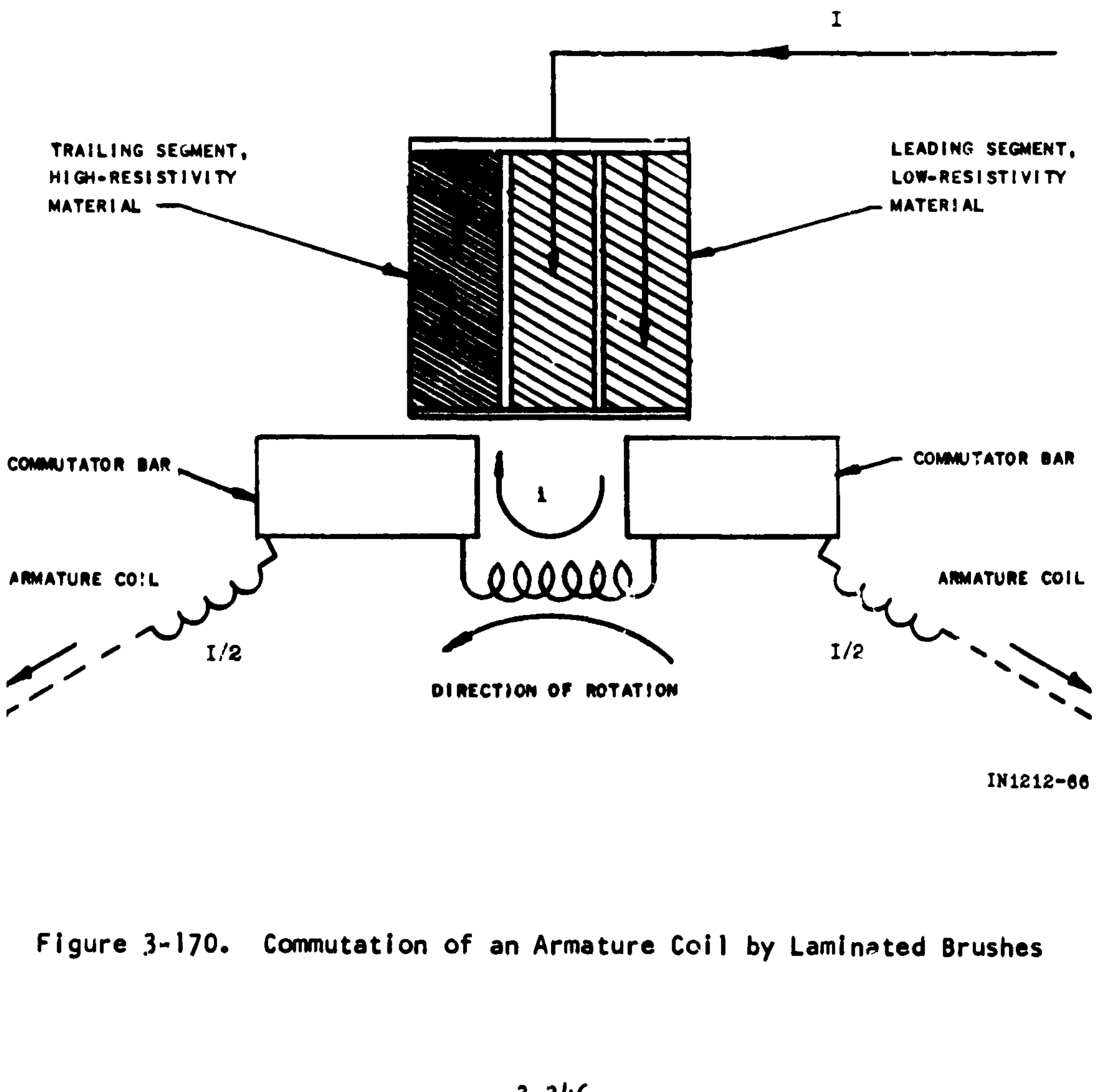


Figure 3-170. Commutation of an Armature Coil by Laminated Brushes

Having the successive segments of the brush increase in resistance, avoids the sharp current drop after the brush leaves the commutator segment. A more linear coil-current reversal results, thus reducing the break transients. The segments of the laminated brush are insulated from one another by some suitable bonding material, and electrically connected by the brush lead or brush spring. Circulating currents, resulting from the self-inductance of the coil under commutation and from the coil-cutting fringe flux from the pole pieces, must flow through the entire length of two brush laminations. The total resistance of this length is much greater than that presented by a direct path across the face of the brush (as would occur with a solid brush). Circulating currents are, therefore, reduced early in the commutation period, and desirable division of current through the two adjacent commutator bars is achieved.

d. Good commutation can be achieved over a fairly wide range of brush positions, relative to the magnetic neutral, so that brush positioning becomes less critical and less dependent upon armature current. The design of laminated brushes should include two or, at most, three laminations. The following criteria should be incorporated in the design:

1) The thickness of the leading-edge lamination of a two-lamination brush should be about 90 per cent of the total thickness, and its resistivity should be as high as allowable for heat dissipation

2) The resistivity of the trailing-edge lamination should be about 15 times that of the leading edge; this lamination should be thick enough to preclude mechanical weakness

3) A thermosetting cement of six-mil thickness should be sufficient to provide electrical insulation between the sections. A cement, that will preclude the formation of a smear of conducting particles from brush wear on the rubbing edge, should be used; it should have a wear-rate equal to that of the brush

4) A brush, with varying resistance characteristics from the leading edge to the trailing edge, can be manufactured without the use of insulating separators and will act somewhat like a laminated brush

e. A copper commutator, in contact with a carbon or graphite brush, develops, after several hours, a layer of copper oxide, mixed with carbon particles, from brush wear. This copper-oxide film introduces unidirectional electrical properties (polarity effects), as in a copper-oxide rectifier. The oxide layer has a nonlinear resistance of higher value at the brush used as cathode, than at the one used as anode. The cathode brush, as a result, passes current in discontinuous, high current-density surges, which cause interference. Approximately ten times as much interference may result from the cathode brush as from the anode brush. Plating the copper commutator with chromium to a thickness of about one mil will reduce the interference level from a cathode brush to that of a relatively quiet anode. No adverse effects will result from the plating; in fact, the hard chromium surface prevents threading and grooving of the commutator. Wear-rate and sliding friction of many brush materials on chromium are of the same order of magnitude as those for copper.

f. Design features that improve commutation also reduce interference generation. The design engineer, faced with using an interference-producing motor or generator, can incorporate several interference-reducing techniques. The most effective and economical technique is the installation of capacitors at the brushes. In generators, for example, installing capacitors at the brushes applies the remedy as close to the interference source as possible. The interference, generated by the commutator and the brushes, will be bypassed to the generated housing. The lead from the brush to the capacitor should be as short as possible, and the capacitor should be bonded to the generator housing to provide a low-impedance path to ground for the interference currents. A good value for such a capacitor is 0.1 μfd with a double voltage rating, depending on the individual machine. Because of the combined interference-generating characteristics of the commutator and the brushes in a dc generator, an additional capacitor is installed at the output (armature) terminal. The preferred installation is a feed-through capacitor through the generator housing. The alternate installation is a 0.1 μf bypass capacitor, mounted externally, to maintain electrical contact with the generator housing and minimize the lead length between the terminal and the capacitor. Figure 3-171 illustrates the mounting of a bypass capacitor at the armature terminal. The installation of capacitors reduces the interference appearing externally on the armature, field terminals, and wiring.

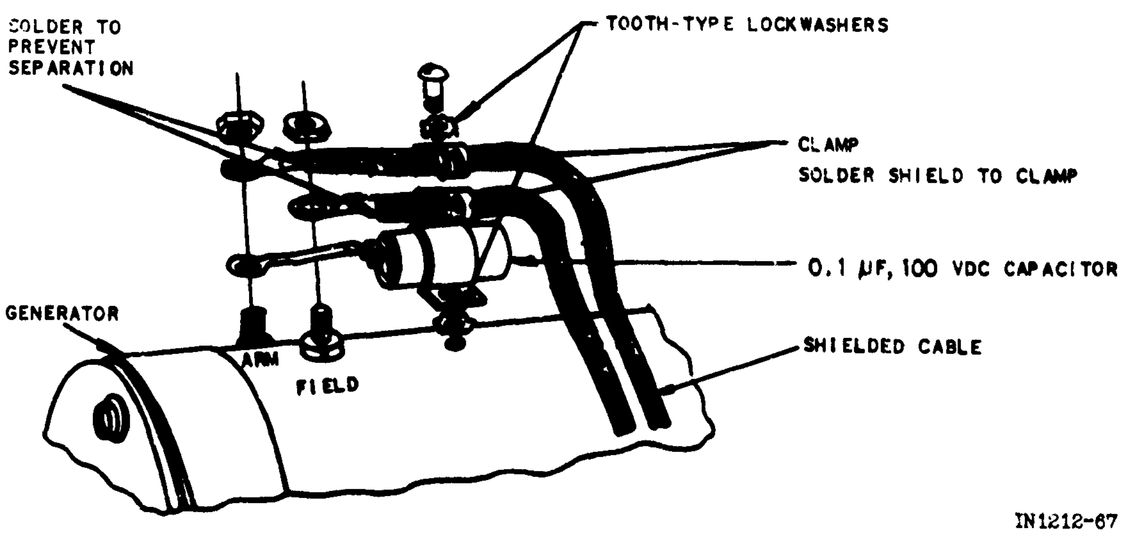


Figure 3-171. Mounting of Bypass Capacitor at Armature Terminal in a DC Generator

g. Over-all shielding is necessary to prevent the radiation of interference from within the generator. This shielding is afforded by the generator housing, which should be designed to provide maximum shielding effectiveness. Ventilation openings should be screened to prevent radiation of interference into space. No matter how perfectly a generator shield is designed, the shaft provides an exit path for interference because it must penetrate shielding. The interference should be bypassed directly to the generator housing by grounding the shaft through a brush, riding on a special grounding slip ring (or riding directly on the shaft). This grounding will also eliminate bearing interference (bearing static or shaft current). Bearing interference results from a periodic discharge of static electricity that takes place, through the bearing, between the shaft and the housing. Eddy currents, induced in the shaft and the housing by the flux lines in the motor, can cause currents to flow through the bearing. These currents can also be caused by certain combinations of armature segments per pole, air gap and permeability inequalities, rotor eccentricities, insulation leakage, or stray electric fields. Another possible source of leakage from the shield is the inspection-band. This band is disadvantageously placed because of its proximity to the interference-generating brushes and commutator; however, its function, of permitting inspection of the brushes and commutator, prevents its being moved to another location. To prevent leakage, the inspection-band should be machined as closely as possible, and should be wide enough to cover the inspection opening adequately, with sufficient overlap to ensure good contact. The band should have bolts, spaced every two inches, to permit secure tightening. Interference gasketing should be installed around the periphery of the opening. After removal of an inspection-band, all contact surfaces on the band and the generator should be thoroughly cleaned before the band is put back into position.

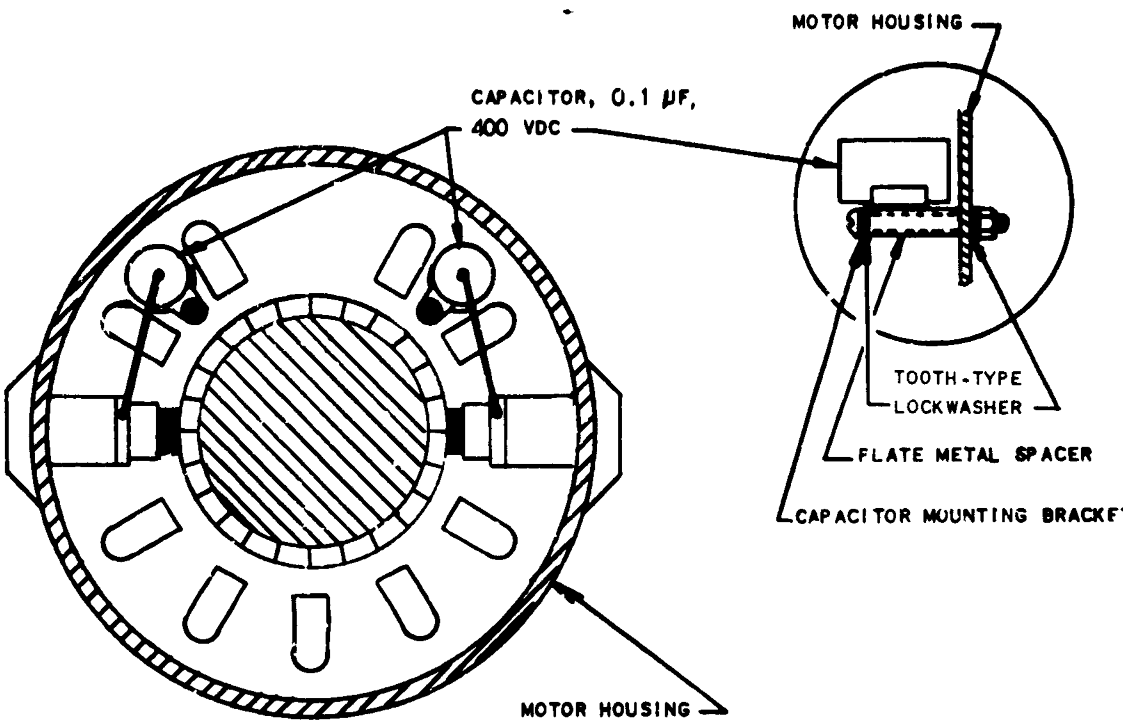
h. The last shielding consideration for a generator housing is to ensure good contacts and low-impedance paths between the three sections of the generator; the two end-plates, and the main housing. This is accomplished by the bonding and shielding practices discussed in Sections III and IV of Chapter 2. The design considerations, for minimizing interference generated by brushes and commutation action in dc generators, also apply to dc motors. Capacitors, installed at the brushes, bypass the generated interference to ground close to the source, providing an effective and economical means of suppression (fig. 3-172).

i. On some dc motors, an adjustable speed control is included in which the field leads are connected to an externally-mounted rheostat. This arrangement necessitates breaking the shield continuity, and therefore enables interference, generated inside the motor, to be conducted out of the housing. Capacitors, installed inside the motor housing connected to these leads, however, will bypass such interference to ground. Figure 3-173 shows a motor with four installed capacitors: one each for the two brushes, and one each for the two field leads. The feed-through capacitor should be mounted at the positive lead (A, fig. 3-174). A less acceptable interference reduction technique for the same motor utilizes bypass capacitors at the brushes (B, fig. 3-174).

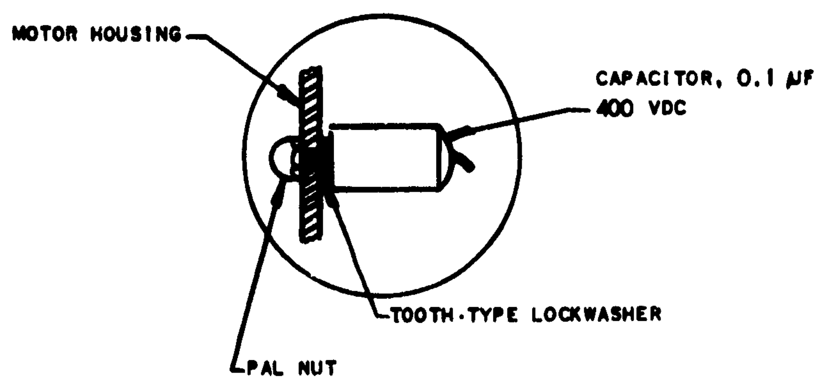
3-60. Alternators and Synchronous Motors

a. Alternators and synchronous motors are very similar to dc generators and motors, except that they supply or use ac, and therefore have slip rings instead of commutators. Commutator interference is absent in these machines. There is, however, interference from the brushes and from the generation of harmonics. Brush interference is lessened because most alternators and synchronous motors have stationary armature and rotating fields; heavy power currents need not be supplied to the rotor. Only the much smaller field currents have to be supplied through the brushes. Because commutation need not be considered in the selection of brushes, the design engineer is permitted a much wider choice in brush pressure, size, and material.

b. In ac generators, careful attention by the design engineer will minimize the generation of harmonics and the resonant conditions that create interference. Production of as pure a sine wave as possible (an



A. GENERAL MOUNTING METHOD FOR CAPACITORS.



B. ALTERNATE MOUNTING FOR CAPACITORS. IN1212-68

Figure 3-172. Capacitor Installation at Brushes in a DC Motor

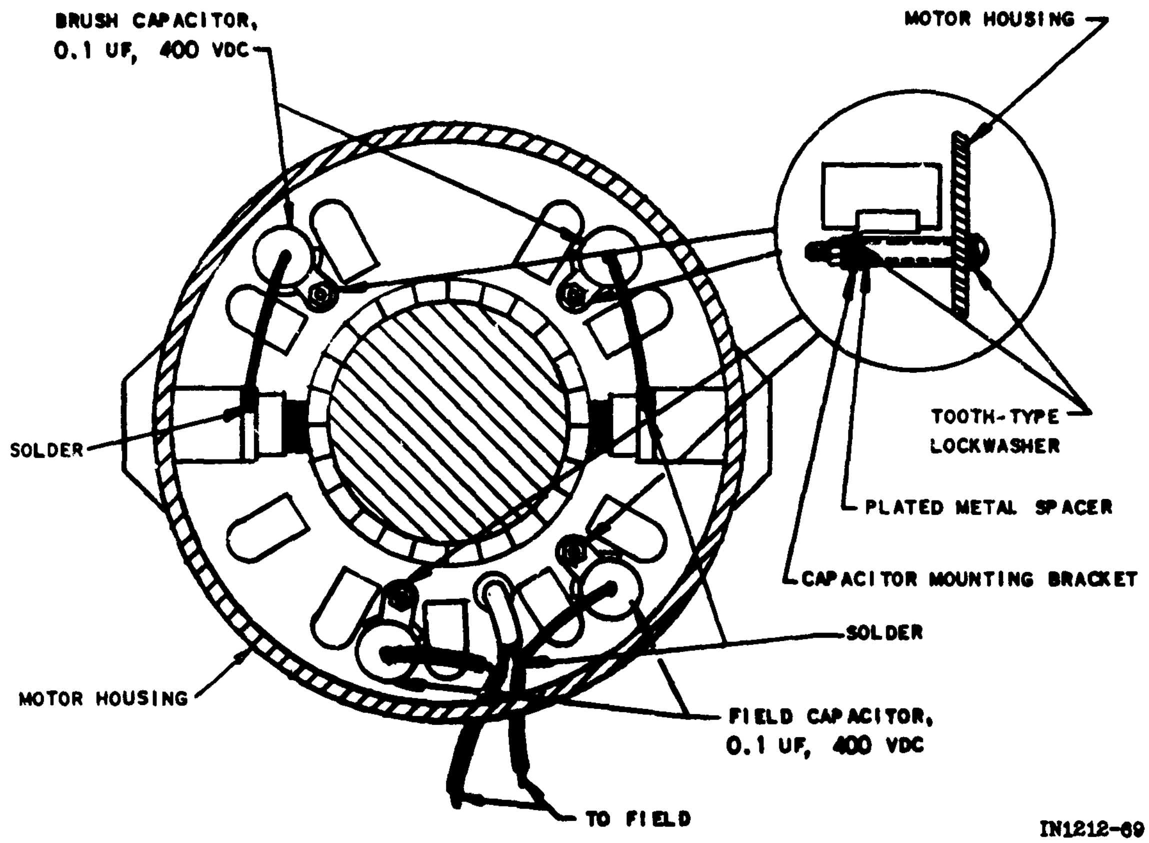


Figure 3-173. Capacitor Installation at Brushes and Field Leads in a DC Motor

important consideration in the design of alternators) is especially important when interference-reduction techniques are considered. A comparatively minute harmonic content may be quite tolerable from all points of view except that of interference reduction. In the reduction of harmonics, special attention must be given to:

(1) Flux distribution. The most important factor determining the waveform of the generated voltage is the distribution of the magnetic flux around the periphery of the armature. Sinusoidal distribution, which produces the least amount of interfer-

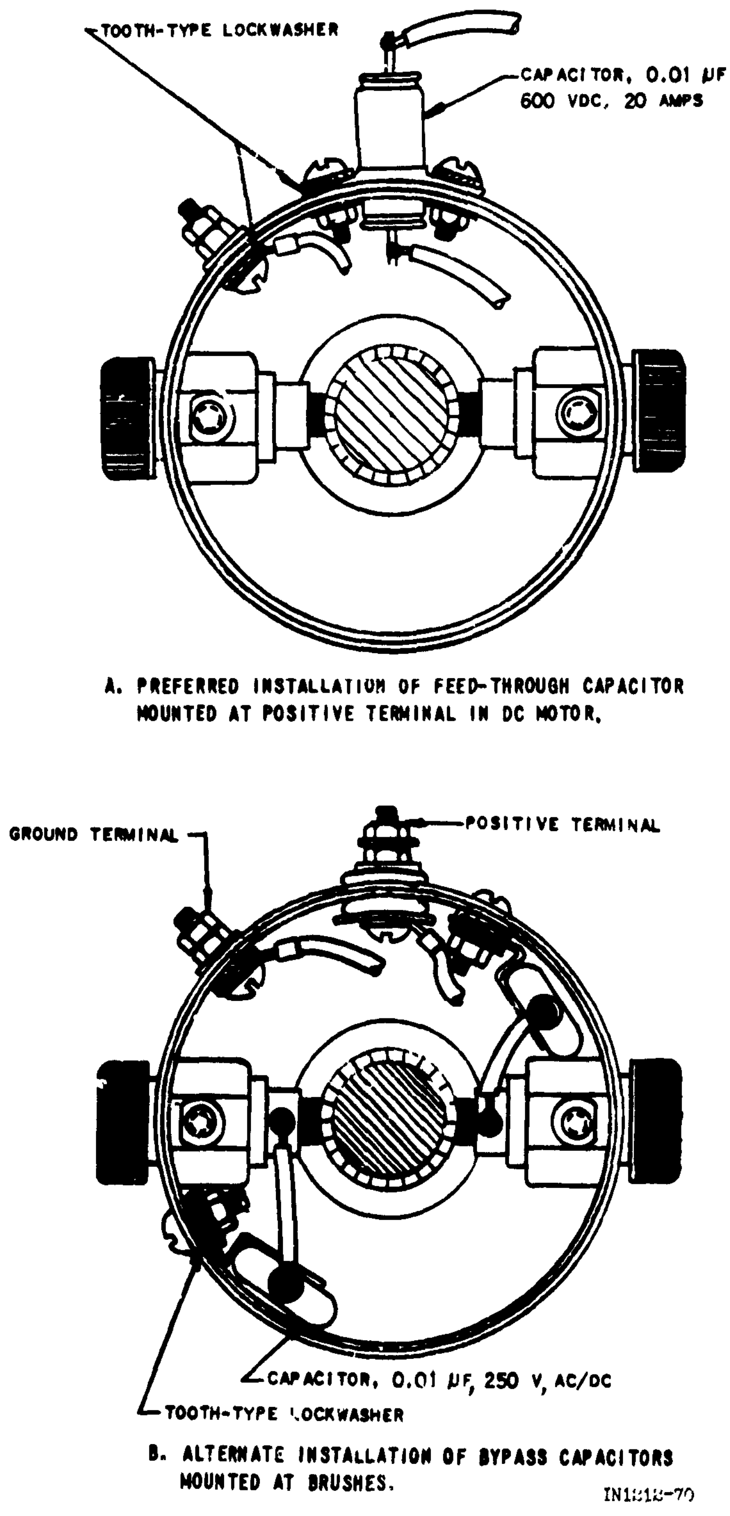


Figure 3-174. Capacitor Installation in DC Motor

ence, may be achieved by chamfering the pole tips or skewing the pole faces.

(2) Symmetry. In a perfectly symmetrical machine, all even harmonics automatically disappear; therefore, special care must be exercised to construct identical pole pieces, to make the yoke and armature perfectly symmetrical, to produce a perfectly uniform winding on the armature, and to avoid all other irregularities.

(3) External connections. In a three-phase alternator, the third harmonic and its multiples disappear at the terminals except when the machine is wye connected and has its neutral grounded, in which case third harmonics are present in the voltage between any phase and neutral. This connection should be avoided, or, if it must be used, special attention should be given to the prevention of the third harmonic and its multiples.

(4) Distribution factor. The distribution factor should be chosen to eliminate the lowest harmonic not eliminated by any of the devices mentioned in (2) or (3).

(5) Tooth ripples. The generation of tooth ripples is greatly decreased by skewing, through one slot pitch, either the pole shoes or the armature slots. Tooth ripples may be eliminated altogether by making the number of armature-slots per pole-pair an odd number. The chord factors, for the harmonics that are contained in the tooth ripples, are then reduced to zero. Slip ring and brush materials should be such that interference is minimized. The design considerations, applied to brushes and commutator surface materials in dc machines, apply equally well here. The effects of brush bounce, due to vibration or irregularities of armature motion, can be minimized by the use of two or more brushes per slip ring.

c. In addition to the interference generated as a result of the brush action on the slip rings and the harmonics present in the sine-wave output of an alternator, the exciter is a prolific source of interference. Because both the exciter (essentially a dc generator) and the ac generator are installed in a single housing, shielding considerations become a combination problem. Plating of the commutator, the use of proper brushes and brush pressure, and the application of bypass capacitors are applicable to the exciter; the other design measures can be applied to the exciter as a separate unit. Although individually designed for interference reduction, the alternator and exciter each generate some interference; this residual interference is reduced by shielding and the use of bypass capacitors installed at terminal outlets.

d. Shielding of the alternator is incorporated in the design of its housing. Low-impedance paths between sections of the housing, provisions for bonding, and screening of all ventilating louvres must be carefully observed if the over-all interference-reduction design is to be effective. As in dc generators, no matter how perfect the shield, a means of escape from the shield for the interference currents is provided by the alternator shaft which penetrates the shield. The same procedures for shielding dc generators therefore apply to alternators. The alternator terminal outlets provide another means of leakage from the alternator. They are prevented from radiating interference by the installation of capacitors. Bypass capacitors are installed inside the terminal strip and are connected to the terminal outlet just before the terminal breaks the shield. This arrangement removes interference from the lead at the last possible point, preventing interference from coupling back into the lead and radiating from the terminals or from their connected wiring. Another type of installation is to mount feed-through capacitors through the terminal strip.

e. The problems of interference suppression for alternators also apply to synchronous motors. Synchronous motors have the same basic components as alternators. A synchronous motor will operate as an alternator, and vice versa. An induction motor should be used instead of a synchronous motor whenever possible because of the lower interference generated by induction motors.

f. The primary source of interference within a single-phase induction motor is the starting device. The starting winding is in series with a switch (or capacitor and switch) that is closed when power is off. When the motor reaches approximately 80 per-cent of its rated speed, the switch is opened (either by centrifugal force or by a solenoid coil) and a single pulse of interference is generated. This switch should be placed in a shielded housing; the leads, leaving the housing, should be filtered.

3-61. Portable Fractional-Horsepower Machines

Portable fractional-horsepower machines include such equipment as portable electric drills and saws. Power is furnished by high-speed, lightweight, ac-dc, or ac electric motors. Such equipment, using ac-dc motors (universal motors), is a major source of interference because commutation is essential in its operation. As in dc motors, an effective, economical method of designing for reduced commutator-brush interference is by installing capacitors at the brushes. In some portable ac-dc machines, restrictions of size and shape prevent the installation of capacitors at the brushes, and it is more feasible and economical to mount the capacitors in other parts of the equipment. Installing capacitors at the line side of the switch bypasses interference to the unit housing at the last point of exit to the power lines, and prevents the interference from coupling back into an interference-free lead, and from being conducted by the power lines. If the mechanical design of the unit prevents the installation of capacitors on the line side of the switch, it is permissible to install them on the motor side. Shielding may be used to ensure that no interference couples back into the leads before they leave the unit.

3-62. Special-Purpose Machines

a. Special-purpose rotating machinery include a variety of equipment; the most important of these are rotary inverters, dynamotors, motor generators, and generators for electric arc-welding equipment. The function of conversion is common to most of this equipment: ac is converted to dc, or to higher frequency ac; or dc is converted to higher or lower voltage dc, or to ac.

b. A rotary inverter, which converts dc to ac, is basically a dc motor with added taps on the armature winding; slip rings are connected to these taps to provide the ac output. Interference is generated by both the ac and dc functions; commutator and brush action in the motor, and brush action and harmonics in the alternator. Figure 3-175 illustrates an interference-reduction design technique for an inverter. The schematic diagram shows two feed-through capacitors bypassing interference from the output leads of the alternator. The dc lead is shielded from the motor by a feed-through capacitor. In addition to the shielding and the feed-through capacitor on the dc line, a capacitor shield is installed to prevent radiation from the terminal on the hot side of the capacitor. This shield also provides a ground for the braid shielding. The ac output leads do not require shielding because the interference generated by the alternator is much less severe than that generated by the dc motor. Bypass capacitors, connected to the brushes in both the motor and the alternator, should be included in the original design. The housing must adequately shield the unit with a feed-through capacitor, mounted through the shield for connection to the dc input lead. The ac leads may not require suppression in addition to that provided by the capacitors at the brushes.

c. A dynamotor (a combination dc motor and generator with a single magnetic field) has an armature with two separate windings and two separate commutators, one at each end of the armature. It transforms low-voltage dc to high-voltage dc, or vice versa. The two commutators make

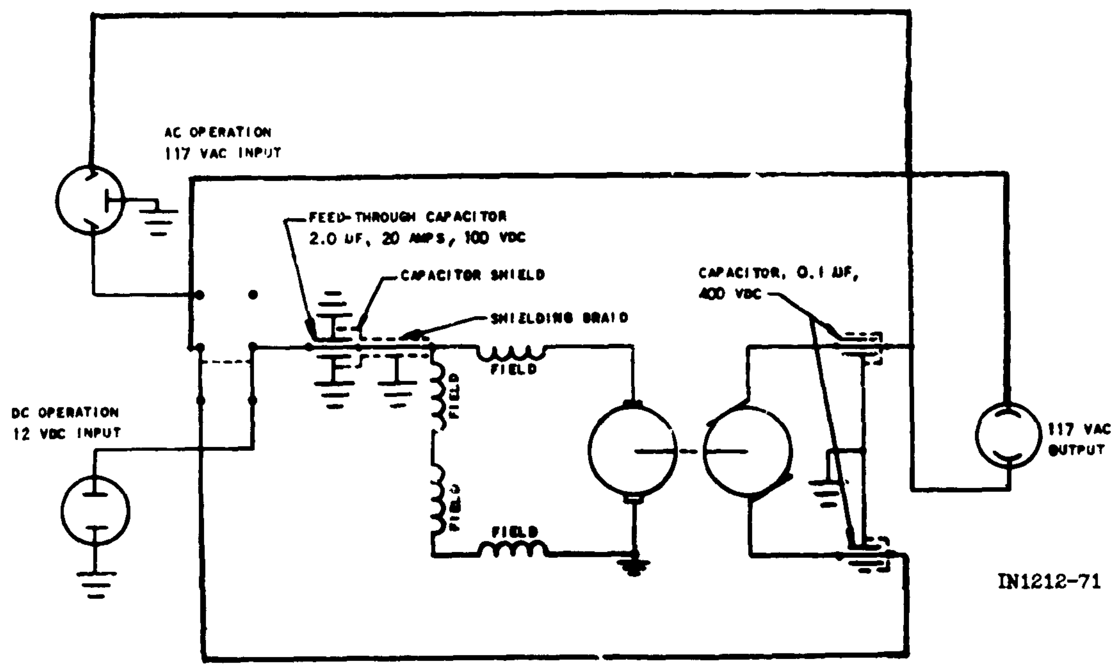


Figure 3-175. Interference Reduction Design Technique for Rotary Inverter

this machine a particularly prolific source of interference. The suppression techniques for dc generators and motors apply to the dynamotor. Figure 3-176 illustrates a dynamotor, with feed-through capacitors bypassing interference to the housing on both the input and output leads. Complete shielding of the dynamotor prevents interference from leaking through other paths.

d. The use of ac commutator motors should be avoided whenever possible. Universal motors fall into this category, as well as repulsion motors and series ac motors. The performance advantage of these types is their high-starting torque; their interference generation, however, is much

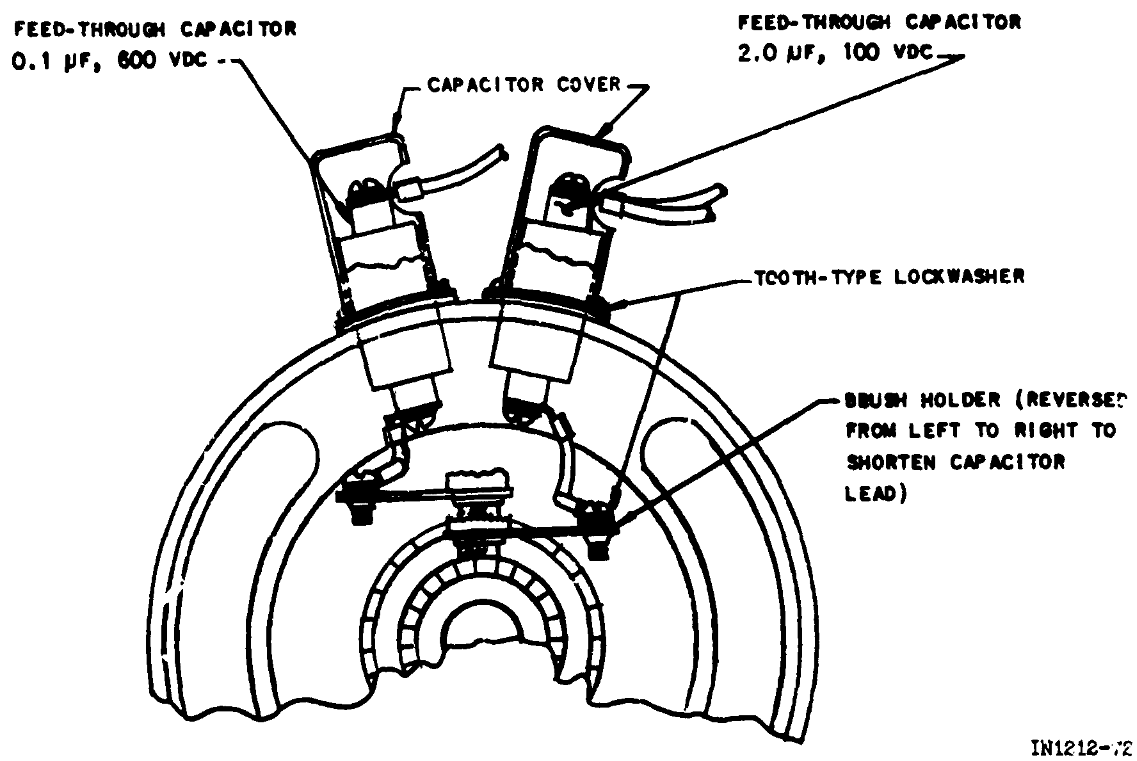


Figure 3-176. Interference Reduction Design Technique for a Dynamotor

more severe than that from other types of ac motors.

e. High-starting torque with ac motors can be obtained without increasing interference generation by using capacitor-type starting, induction-run motors. These motors use a high-capacitance condenser for starting purposes only. Starting torques of 200 to 350 per-cent of full-load torque are feasible with acceptable starting currents. Ratings from 1/8 to 10 hp are available.

f. Generators, for electric-arc equipment, require special attention only when connected to such a severe source of interference as the electric arc. The generator can be either ac or dc, and be driven either by an ac or dc motor or an engine. Nothing can be done to reduce the interference generated by the electric arc itself. The equipment should be located away from communication equipment, and in buildings with good shielding characteristics. The leads, from the generator to the welding electrodes, can become very effective interference radiators and should be adequately shielded.

3-63. Summary of Interference-Reduction Design Techniques for Rotating Machinery

a. DC Generators.

(1) Install capacitors at the brushes.

(2) Install a capacitor at the armature terminal, either feed-through or bypass.

(3) Shield the housing. The housing should have screened louvres, conducting gaskets between sections, and, if an inspection plate is needed, it should be tight-fitting and gasketed.

(4) Shield the terminals. They should be covered with individual caps that terminate in threaded fittings, or with a shield, covering both terminals which terminate in one threaded fitting.

(5) Shield the interconnecting wiring between the voltage regulator and the generator.

(6) Install a shaft bond.

(7) Maintain good bonding between the generator and the driving engine.

b. DC Motors.

(1) Install a feed-through capacitor at the positive terminal to eliminate the need for capacitors at the brushes and shielding

of external wiring.

(2) Install capacitors at the brushes (alternate installation to item 1).

(3) If the motor is equipped with an adjustable speed control, install capacitors inside the housing at the field leads. The field leads are bypassed just prior to their exit from the motor housing.

(4) Shield the housing. The housing should have screened louvres and conducting gaskets between sections.

(5) Maintain good bonding between the motor and ground through either direct mounting or through the use of bond straps.

c. Alternators.

(1) Install capacitors at the slip-ring brushes.

(2) Install capacitors at the exciter brushes.

(3) Mount feed-through capacitors through the terminal strip in the output leads, or install bypass capacitors inside the terminal strip and connect them to the terminal outlet just before the terminal breaks the shield; this makes shielding of the alternator output leads unnecessary.

(4) Shield the housing. The housing should have screened louvres and conducting gaskets between sections.

(5) Install a shaft bond.

(6) Install braided shielding on the lead from the exciter to the voltage regulator.

d. Synchronous and Induction Motors. The same measures that apply to alternators apply to synchronous motors. An induction motor, however, should be used instead of a synchronous motor whenever possible because it generates less interference.

e. Portable Fractional-Horsepower Machines with Universal Motors.

(1) Install capacitors at the brushes.

(2) Install capacitors on the line side of the switch at the last possible exit of interference onto the power lines.

(3) Shield the unit housing.

f. Rotary Inverters.

(1) Install capacitors at both the commutator and slip-ring brushes. The ac output leads ordinarily do not require shielding if capacitors are installed at the slip-ring brushes.

(2) Shield the unit housing.

(3) Mount a feed-through capacitor through the shield for connection to the dc input lead.

g. Dynamotors. The same techniques apply for dynamotors as for dc generators and motors.

h. Generators for Electric-Arc Equipment. Leads from the generator to the welding electrodes should be shielded.

BIBLIOGRAPHY

BIBLIOGRAPHY

The basic directive relating to interference reduction is: Memo for the Secretaries of the Army, Navy and Air Force, subject "Radio Frequency Compatibility", dated July 19, 1960, with attachment "Radio Frequency Compatibility Program" dated 9 June 1960.

In addition to the material listed here, excellent general interference bibliographies have been prepared by Georgia Institute of Technology under Contracts DA36-039 sc-74855, DA36-039 sc-87183 and DA36-039 AMC-02294(E).

Listed in the following bibliography are articles, papers, reports and texts to which the reader may refer for more discussion pertaining to specific areas covered in this Guide. All bibliography entries are coded to indicate security classification as follows:

(U) Unclassified
(C) Confidential
(S) Secret

1. Absorbers for Electro-Explosive Devices, Development of Broad Band Electromagnetic (C) Monthly Progress Report, RF Attenuating Material Studies, The Franklin Institute, Mar. '60

2. Absorbing Materials, Broadband (U) W. H. Emerson, A. G. Sands, M. V. McDowell, Tele-Tech & Electronic Industries, Nov. '55

3. Adjacent Channel Interference from Frequency Shift Keyed Carriers, Reduction of (U) U.S. Department of Commerce National Bureau of Standards, June 25 '57

4. Airborne Radio Frequency Interference Problems, Analysis of (U) Filtron Company, Inc., April 30 '62

5. Aircraft, A General MIL Test Plan for KC-135 (C) Filtron Company, Inc., Aug. 5 '60

6. Aircraft Altimeters, Determination of Causes & Extent of Interference from (S) Final Report Filtron Company, Inc., Feb. 23 '62

7. Aircraft Interface Study, AN/USD-7 (S) Cutler Hammer, Feb. 9 '62

BIBLIOGRAPHY

8. Aircraft Interface Study, Addendum to AN/USD-7 (S) Filtron Company, Inc., May 17 '62

9. Aircraft, RF Interference Test of Hughes (C) Filtron Company, Inc., Mar. 16 '59

10. ALQ-35 System Tech. Report, Radio Interference Test on Emerson (C) Filtron Comapny, Inc., Nov. 3 '59

11. Altimeters, Determination of Causes and Extent of Interference to BMEWS from Aircraft, Final Report, (S) Filtron Company, Inc., Feb. 23 '62

12. Aluminum Foil Tape for Application Over Coaxial Cables, Evaluation of (C) Filtron Company, Inc., July 11 '56

13. Amplifier, RFI Test Report on Band 10 RF Distribution (S) Filtron Company, Inc., March 2 '62

14. AMQ-15 Storm and CB&T Radars, Interference Considerations for (U) Filtron Company, Inc., March 12 '59

15. AN/USD-7 Band 8, RFI Test Report on (C) Filtron Company, Inc., June 20 '62

16. Arc Producing Devices, Preliminary Draft (U) Jansky & Bailey, Inc., June '60

17. Arc-Suppression Network, Three-Component (U) Electronics, Sept. 2 '60

18. Arc Welding Equipment, Control and Suppression of Radio Interference from High Frequency Stabilized (U) Canadian Dept. of Transport, Telecommunications Div., June 3 '53

19. Arrays, Specification of (U) Military Systems Design, May-June '62

20. Attenuators, S- and L- Band Ferrite, Final Report, (U) Sperry Microwave Electronics Company, May '60

21. Beacon, Radio Interference Evaluation of Range Safety (C) Filtron Company, Inc., Sept. 26 '56

22. Blanker Model RB-128, Engineering Data for Self-Contained Radar Interference (U), Empire Devices Products Corp.

BIBLIOGRAPHY

23. BMEWS, Determination of Causes & Extent of Interference to, Final Report (S) Filtron Company, Inc., Feb. 23 '62

24. Bomarc Missile Test Equipment, Design Guidance for Interference-Free Operation of, Phase I (U) Frederick Research Corporation, Mar. 29 '56

25. Bonding, Grounding and Cable Selection, An Integrated Approach to (U) Filtron Company, Inc.

26. Bonding Materials - Metallic Mating Surfaces: Low RF Impedance, Final Report, (U) Armour Research Foundation, Nov. '54

27. Bonding Requirements for Interference Reduction, Evaluation of Radio Frequency (S) Filtron Company, Inc., Nov. 11 '58

28. Cable and Splice, Electromagnetic Shielding Efficiency Tests for Western Electric Lepath, Section 10A (U) Filtron Company Inc., April '59

29. Cabling Considerations, General (U) Filtron Company, Inc., June 17 '58

30. Call-7 System Radio Frequency Program, First Monthly Report for (S) Filtron Company, Inc., Oct. 31 '60

31. Capacitor, The C-D, Vol. 4, No. 8 (U) Cornell-Dubilier Electric Corp., Nov.-Dec. '40

32. Capacitor Types, Radio Interference Suppression Characteristics of Various (U) Scott L. Shive, Technical Memorandum No. M-1278, U.S. Army Electronics Lab., Ft. Monmouth, N.J.

33. Capacitors, Cascaded Feedthrough (U) H. M. Schlicke, Proceedings of the IRE, May '56

34. Capacitors, How to Select Optimum Values for RF ByPass (U) Dr. Richard E. Lee, Electronic Design, May 27 '59

35. Capacitors, Insertion Loss Measurements of Feedthrough (U) Filtron Company, Inc., Sept. 21 '59

36. Capacitors, Short Leads of (U) Canadian Dept of Transport, Radio Division, Suppression of Inductive Interference, Jan. 7 '43

37. Circuits and Tubes, Electronic (U) Cruft Laboratories, Harvard University, 1947 McGraw-Hill, Pub.

BIBLIOGRAPHY

38. Co-Channel Interference Study, Report on (C)
Filtron Company, Inc., Oct. '56

39. Collection and Analysis of Interference Applied Research Data (U)
Filtron Company, Inc., Nov. 7 '60

40. Communications Electronics Interference, Location, Identification, Suppression of (U) Standard Practices Manual, Nov. 1 '57, USAF T. O. 31-1-48

41. Communications - Electronics Systems, Engineering Interference-Free (U) Frederick Research Corp., Standard Planning Manual Jan. 15 '58

42. Communication Interference Reduction Techniques, Investigation and Study of (U)
ASTIA, Document AD-148844

43. Communications Interference in the 1960-62 Pentomic Field Army, Project Monmouth IV (C) U.S. Army Signal Research & Development Lab, May 1 '59

44. Communications Systems Study, Project Monmouth III, Annex E (C)
U.S. Army Signal Research & Development Lab, Oct. 8 '58

45. Commutators and Brushes, Suppression of Inductive Interference Originating at (U) Canadian Dept of Transport, Telecommunications Div., Jan. 5 '39

46. Compatibility, Electromagnetic - Lecture Series (U)
Armour Research Foundation, Aug.-Sept. '61

47. Computer Design, Shielding in Modern (U)
Automatic Control, Dec. '58

48. Conducted Electromagnetic Interference Measurement Survey 500 - 130,000 Volts, Final Report (U) Westinghouse Research Laboratories, Dec. 31 '61

49. Conductors, Radio Interference of Bundled (U)
G. E. Adams, Electrical Engineering, Jan. '57

50. Conference on Radio Interference Reduction, Proceedings of the (U)
Armour Research Foundation, Dec. '54

51. Countermeasures Analysis and Identification Manual, Vol II, Electronic (C) Bureau of Naval Personnel, Dec. 27 '56

BIBLIOGRAPHY

52. Contermaasures Search & Analysis Techniques, Final Report 44-F (S) Haller, Raymond and Brown, Inc., Apr. 30 '54

53. Control Program for Weapon System TF-X, Electrical and Electromagnetic Interference - Technical Proposal (S) Filtron Company, Inc., March 22 '61

54. Corona Effects, Analysis & Elimination of (U) C. E. Mercier, W. E. Elliott, Electrical Manufacturing, Nov. '57

55. Corona Tests, Scope and Instrumentation, High Altitude (U) American Institute of Electrical Engineers, Jan. '57

56. Decibels and Nepers - Definitions and Calculations (U) J. Zdzieborski and R. E. Post, Electrical Design News, July '59

57. Definitions and Measurements, Radio Interference (U) Radio Corporation of America

58. Design Guide, Project Engineers (U) U.S. Army Signal Research and Development Lab, July 2 '62

59. Doors, Design Review RF Lock Shielded (U) Filtron Company, Inc., March '60

60. Design Techniques for Interference Free Operation of Airborne Electronic Equipment (U) Wright Patterson AFB (Frederick Research Corp.) 1952

61. Diodes for RF Noise, Testing (U) J. C. Seen (Convair), Electronic Equipment Engineering, Jan. '59

62. Direct Current Supplies, Low-Noise Stabilized (U) Electronic & Radio Engineer, Sept. '57

63. Electric Circuits and Fields (U) Harold Pender and S. Reid Warren, Jr., McGraw-Hill, 1943

64. Electric and Magnetic Fields (U) Stephen S. Attwood, John Wiley & Sons, 1949

65. Electrical Apparatus, Suppression of Inductive Interference from Small (U) Canadian Dept of Transport, Oct. '37

66. Electrical Interference (U) Instruments & Automation, June '58

BIBLIOGRAPHY

67. Electron Tubes in Military Equipment, Techniques for Application of, Vol. II (U) Wright Patterson AFB, Project No. 4156, June '59

68. Electronic Equipment Characteristics Non Radar Types, Directory of, Final Technical Report Project No. A-343 (C) Signal Engineering Lab, Fort Monmouth, July '57 to Sept. '59

69. Electromagnetic Compatibility, Eighth Tri-Service Conference on (U), Armour Research Foundation, Oct. '62

70. Electromagnetic Interference, Proceedings of the Classified Session of the Symposium on (S) U.S. Army Signal R&D Lab, Fort Monmouth, June '58

71. Electromagnetically Shielded Areas, Theoretical and Practical Aspects of Testing Large (U) Filtron Company, Inc., Aug. '61

72. Enclosure Design, Optimum (U) Filtron Company, Inc., 1959

73. Enclosure Openings, Designing Noisefree (U) Arnold Albin, Electronics Engineering, Aug. '58

74. Environment at NASA Missile Test Facilities - Redstone, Alabama, Survey of Radio Interference (S) Filtron Company, Inc., July '60

75. Evaluation & Recommendation, A Preliminary - Sycamore Canyon Site (C) Filtron Company, Inc.

76. Feed-Through Capacitors, Insertion Loss Measurements of, Test Report on (U) Filtron Company, Inc., Sept. '59

77. Field Strength with Distance, Variation of - Inductive Interference (U) Canadian Dept of Transport, Ottawa, Canada, June '54

78. Filters, A Guide to RFI (U) Mervin H. First, Electronic Industries, June '60

79. Filters, Band-Pass (U) Fuad Surial Atiya, Wireless Engineer, Dec. '53

80. Filters Aid Miniaturization, Ceramic (U) Electronic Industries, Nov. '59

81. Filter Design by Impedance Chart (U) L. Goeller, Jr., Electronic Equipment Engineering, March '62

BIBLIOGRAPHY

82. Filter Design, Using Insertion Parameters for (U)
Electronic Industries, Feb. '61

83. Filters, How to Specify (U)
Stanley Boyle, Electronic Industries, Sept. '58

84. Filters, Radio Interference (U)
John W. Trankaus, Sperry Engineering Review, Nov.-Dec. '56

85. Filters Using Ferrite Inductor Cores, Design Methods for Radio Interference (U) Convair, July '57

86. Filters Without Fears (U)
Thomas Roddam, Wireless Work, Dec. '54

87. Flexible Shielding Conduit, Shielding Effectiveness of - Memorandum Report (Restricted) Aircraft Radio Laboratory, Aug. '44

88. Fluorescent Lamps, A Preliminary Study of Radio Interference as Caused by (U) L. F. Shorey, S. M. Gray, Illuminating Engineering

89. Fluorescent Lamps, Shielding of Radiated Radio Interference from (U) Corning Glass Works, Sept. '60

90. Fluorescent Lighting, Suppression of Inductive Interference from (U) Canadian Dept of Transport, Telecommunications Div., Mar. '61

91. Frequency Allocation and Assignment, Project Monmouth II, Summary Report, (S) Signal Corps Engineering Labs, Fort Monmouth, Oct. '56

92. Frequency Assignment for Field Army Communications, A New Method of Predetermining - Project Monmouth IV Study (C)
U.S. Army Signal Research & Development Lab, April '59

93. Frequency Assignments for BMEWS Site II, Final Report (S)
Filtron Company, Inc., May '61

94. Garments for RF Protection of Personnel in High Power Density Environments (S) Filtron Company, Inc., March '59

95. Generators, Suppression of Interference from Large DC (U)
Canadian Dept of Transport, Telecommunications Division

BIBLIOGRAPHY

96. Grounding and bonding Practices (U)
Filtron Company, Inc.

97. Hazards Survey, Nike Radiation (U)
Filtron Company, Inc., April '60

98. Honeycomb Shielding, Insertion Loss of (U)
Filtron Company, Inc., Dec. '60

99. Honeycomb Shielding from 1 to 10 Kilo-Megacycles, Insertion Loss of (U) Filtron Company, Inc., Jan. '61

100. Hughes, Radio Interference Evaluation of (C)
Filtron Company, Inc., Jan. '59 to March '59

101. Inductive Interference, Suppression of (U)
Canadian Dept of Transport, July '54

102. Inductive Interference Suppression (U)
Canadian Dept of Transport, Telecommunications Div., Dec. '53

103. Inductively Coupled Traps, Designing (U) William F. Gariano, Tele-Tech & Electronic Industries, Apr. '54

104. Intercept Techniques, Analysis and Long Range, Final Report No. 63F, (C) Haller, Raymond and Brown Inc., Apr. '57

105. Interference, Electrical (U) Dr. James C. Coe, Instruments and Automation, June '58

106. Interference, Navy Manual on (U)
Navy Dept., Code No. 837

107. Interference, Proceedings of the Unclassified Sessions of Symposium on Electromagnetic (U) U.S. Army Signal & Research Lab, Fort Monmouth, June '58

108. Interference Monitoring Study Status Report (S)
Filtron Company, Inc., April '59

109. Interference Reduction Techniques, Evaluation of (U)
The Electro-Mechanics Co., April '58

110. Intersystem Interference Study, Down Range, Vol A (S)
Filtron Company, Inc., May '61

111. Intersystem Interference Study Appendixes, Down Range (S)
Filtron Company, Inc., May '61

BIBLIOGRAPHY

112. Jammer Location Systems, Armed Services Technical Information Agency - An Evaluation of (S) Ramo-Wooldrige, Feb. '59

113. Lighting Plants, Suppression of Inductive Interference from Small (U) Canadian Dept of Transport, Telecommunications Division, Feb. '52

114. Measurement of Radio Interference, Preliminary Work by the British Standards Institution Committee on the (U) American Standards Assoc., Nov. '44

115. Measurements, Electromagnetic Field Theory as Applied to Radio Interference (U) William Jarva, Filtron Company, Inc.

116. Measurements, Technical Note Problems of Radio Interference (U) U.S. Naval Air Development Center, Feb. '55

117. Measurements on Ferrite Cores, RF (U) Pietro Lombardini, Richard F. Schwartz, Electrical Manufacturing, April '59

118. Memory Unit Model 3100, Radio Interference Test Report on Rese (U) Filtron Company, Inc., Aug. '61

119. Mercury Arc Rectifiers, Suppression of Inductive Interference from (U) Canadian Dept of Transport, Telecommunications Div., Jan. '39

120. Metal Seam, Effect of Holes on the Shielding Efficiency of a Sprayed (U) Filtron Company, Inc., May '58

121. Microelectronics, Progress and Pitfalls in (U) Electronics, Oct. '62

122. Microwave Communication Services, Radar Interference to (U) R. D. Campbell, American Institute of Electrical Engineering, Apr. '58

123. Microwave Filters, Tem Mode (U) D. V. Geppert, R. H. Koontz, Tele-Tech & Electronic Industries, Nov. '55

124. Microwave Interference, Measuring (U) Electronic Design, Oct. '58

125. Microwave Interference, Measuring (U) Robert Saul, Electronic Design, Nov. '58

126. Microwave RFI Instrumentation, A Recent Advance (U) Stoddart Aircraft Radio Company, Oct. '62

BIBLIOGRAPHY

127. Microwave Systems, Frequency Plan for 6 GC (U) B. R. Hallford American Institute of Electrical Engineering, Dec. '61

128. Missile, Compilation Data Obtained During Radio Interference Evaluation of Hughes Aircraft GAR-3 (C) Filtron Company, Inc., Mar. '59

129. Missile Electrical and Electronic Systems, Reduction of Radio Interference in (U) Filtron Company, Inc., July '57

130. Missile Electrical and Electronic Systems, Reduction of Radio Interference in (U) Filtron Company, Inc., Sept. '57

131. Missile, RF Interference Test of Hughes Aircraft Co. GAR-3 (C) Filtron Company, Inc., Dec. '58

132. Missile Telemetering System, Analysis of Interference in a, Report (U) A. Albin, L. Milton, Filtron Company, Inc., June '58

133. Mixer Design Charts, Sum and Difference (U) Electronics, Dec. '59

134. Monitor, Radio Interference Investigation of Broadband Noise Appearing on the Range Safety Command (C) Filtron Company, Inc., Nov. '56

135. Monitoring and Alarm System for BMEWS, Design of a Radiation, Final Report (C) Filtron Company, Inc., March '59

136. Monmouth II, Project - Summer Map Study (C) Filtron Company, Inc., Jan. '56

137. Monmouth II Monthly Progress Report (C) Filtron Company, Inc., April '56

138. Motors and Generators, Suppression of Inductive Interference from Commutators of (U) Canadian Dept of Transport, March '49

139. Naval Equipment, Radio Interference Suppression Techniques for (U) International Electronics Engineering, Inc., July '55

140. Navigation and Pod Guidance System Installation, Radio Interference Analysis for a Bombing - Special Report (S) Filtron Company, Inc., Sept. '57

141. Navigation and Pod Guidance System, Development of Light Weight Suppression Methods B-58 Bombing - Summary Report (C) Filtron Company, Inc., Nov. '57

BIBLIOGRAPHY

142. Netic and Co-Netic Magnetic Shielding Manual (U)
Perfection Mica Company

143. Noncommunication Systems, Progress Report, Projection Monmouth III Annex D (C) U.S. Army Signal R&D Lab, Fort Monmouth, Oct. '58

144. Norad Conference (S) Ent. AFB, Colorado
Filtron Company, Inc., May '60

145. Oven Card, RFI Test Report on Valpey Crystal (U)
Filtron Company, Inc., June '61

146. Packaging Considerations Necessary for RF Interference Suppression, Mechanical (U) Filtron Company, Inc., May '59

147. Packaging Ideas, Electronic (U)
Electrical Design News, July '59

148. Packaging, Product (U)
J. R. Bathrick, Electronic Products, Dec. '62

149. Parametric Amplifiers for Microwave Communications Systems (U)
J. N. Ratti, American Institute of Electrical Engineering, Jan. '62

150. Pod System, Radio Frequency Interference (U)
Sperry Gyroscope Company, June '56

151. Progress Report, Filtron Project, Third Monthly (S)
Filtron Company, Inc., Nov. '56

152. Project Monmouth II Monthly Progress Report No. 2 (C)
Filtron Company, Inc., Dec. '55

153. Project Monmouth II Monthly Progress Report No. 4 (C)
Filtron Company, Inc., Mar. '56

154. Project Monmouth II Monthly Progress Report No. 6 (C)
Filtron Company, Inc., Apr. & May '56

155. Project Monmouth II Monthly Progress Report No. 7 (C)
Filtron Company, Inc., June '56

156. Project Monmouth III Progress Report, Including Annexes A, B, C (C) U.S. Army Signal R&D Labs, Fort Monmouth, Oct. '58

BIBLIOGRAPHY

157. Project 1124, Selected Reports on (C)
Filtron Company, Inc., Feb. '57

158. Project MX-1964 Final Report (C)
Filtron Company, Inc., Dec. '58

159. Propagation Considerations in RFI (U) L. Valcik, R. B. Schulz,
Electronic Industries, Dec. '60

160. Pygmy Connectors, Electromagnetic Coupling Tests Relating to
Bendix (C) Filtron Company, Inc., Dec. '58

161. Radio Central, Systems Study of the Combat Command (C)
U.S. Signal Corps Engineering Labs, April '55

162. Radio Central, Systems Study of the Command Post Technical
Memorandum M-1662 (C) Signal Corps Engineering Labs, Fort
Monmouth, Apr. '55

163. Radio Frequency Interference is Everybody's Business (U)
John E. Hickey, Jr., Electronic Industries, March '60

164. Radio Frequency Interference Reduction, Lecture Notes on -
Sections I & II (U) Filtron Company, Inc., Oct. '58

165. RFI Problems, Better Designs, Test Methods Combat (U)
Alfred Rosenblatt, Electronic Design, July '62

166. RFI Levels, Finding (U)
James G. Arnold, Electronics, Nov. '59

167. RFI Standard, Proceedings of the IRE (U)
The Institute of Radio Engineers, Inc., Sept. '61

168. Radio Interference, Cause, Effects, and Suppression, An Outline
and Notes on (U) RCA, Engineering Products Division

169. Radio Interference, How to Suppress (U)
Electrical Mfg., Sept. '54

170. RFI, Recent Developments in (U)
IRE Transactions on RFI, May '62

171. Radio Interference Reduction, Proceedings of the Second
Conference on (U) Armour Research Foundation, March '56

BIBLIOGRAPHY

172. Radio Interference Reduction and Electronic Compatibility, Proceedings of the Fourth Conference on (U) Armour Research Foundation, Oct. '58

173. Radio Interference Reduction and Electronic Compatibility, Fifth Conference on (S) Armour Research Foundation

174. Radio Interference Reduction and Electronic Compatibility, Sixth Conference on (U) Armour Research Foundation of Illinois Inst. of Technology, Oct. '60

175. Radio Interference Reduction and Electronic Compatibility, Seventh Conference on (U) Armour Research Foundation, Nov. '61

176. Radio Interference, Reduction of Power-Line (U) U.S. Dept. of Commerce, Office of Technical Services, Feb. '55

177. Radio Interference Research Report, Final, Vol. III (U) International Electronics Engrg., Inc., Washington, D. C.

178. Radio Interference Suppression (U) Depts of the Army and Air Force TM 11-483, Oct. '56

179. Radio Sets, Interference Design Study Tactical, Final Report (C) HRB Singer, Inc., Dec. '60

180. Radio Systems for High Noise Paths, Voice (U) Electronics, Dec. '59

181. Radar and Communications Receivers with Application to BMEWS, Some Considerations of Pulse Type Interference to (S) Filtron Company, Inc., Oct. '58

182. Radar, Design Study of a gh Resolution Reconnaissance (Title U) (S) Goodyear Aircr Corp., Arizona, May '58

183. Radar Receiver Employing a Comb Filter for the Detection of Doppler Information, Adjacent Channel Interference in a (S) Filtron Company, Inc., July '58

184. Radar in Transmitter Building 3, Thule, Greenland, Analysis of Radio Interference Problems Due to Installation of BMEWS Tracking (S) Filtron Company, Inc., Nov. '60

185. Radars, Instrumentation for Spectrum Signature Measurement of Pulse Compression (U) Jansky and Baily, Division of Atlantic Research Corp., Washington, D.C.

BIBLIOGRAPHY

186. Radar Interference, Reducing Mutual (U)
Electronics, July '59

187. Radars, Methods of Interference Reduction in Search (S)
The Rand Corp., Apr. '58

188. Radar 10, Reduction of Radio Frequency Interference in the Sperry (U) Sperry Piedmont Corp. for Sixth Conference on Radio Interference Reduction, Oct. '60

189. Radar's Running Rabbits (U)
Richard A. Wall, Electronics, July '59

190. Radar Set APS 42, Radio Interference Evaluation of, Final Report (U) U.S. Naval Air Development Center, Aug. '51

191. Radiation Hazards Investigation for Radio Corporation of America (U) Filtron Company, Inc., April '58

192. Reconnaissance System, Fifth Monthly Report for the ASD-1 Electronic (S) Filtron Company, Inc., April '59

193. Reconnaissance System, Sixth Monthly Report for the ASD-1 Electronic (S) Filtron Company, Inc., April - May '59

194. Reconnaissance System, Eighth Monthly Report for the ASD-1 Electronic (S) Filtron Company, Inc., June - July '59

195. Reconnaissance, Radio Frequency Interference Control and Suppression of ASD-1 Electronic System (S) Filtron Company, Inc., Mar. '58

196. Reconnaissance System, First Monthly Report for the ASD-1 Electronic (S) Filtron Company, Inc., Nov. '58

197. Reconnaissance System, Third Monthly Report for the ASD-1 Electronic (S) Filtron Company, Inc., Jan. - Feb. '59

198. Reconnaissance System, Final Engineering Report for ASD-1 Electronic (S) Filtron Company, Inc., Mar. '59

199. Reconnaissance System, Ninth Monthly Report for the ASD-1 Electronic (S) Filtron Company, Inc., July - Aug. '59

200. Reconnaissance System, Tenth Monthly Report on ASD-1 Electronic (S) Filtron Company, Inc., Nov. '59

BIBLIOGRAPHY

201. Reconnaissance System, Second Monthly Report for the USD-7 Electronic (S) Filtron Company, Inc., Dec. '58

202. Reconnaissance System, Eleventh Monthly Report for the USD-7 Electronic (S) Filtron Company, Inc., Oct. - Nov. '59

203. Reconnaissance System, Twelfth Monthly Report for the USD-7 Electronic (S) Filtron Company, Inc., Feb. '60

204. Reconnaissance System, Fourteenth Monthly Report for the USD-7 Electronic (S) Filtron Company, Inc., Mar. - Apr. '60

205. Reconnaissance System, Fifteenth Monthly Report for the USD-7 Electronic (S) Filtron Company, Inc., June '60

206. Reconnaissance System, Radio Interference Suppression, AN/USD-7 Electronic, Monthly Technical Progress Report (U) Filtron Company Inc., June '60

207. Reconnaissance System, Seventeenth Monthly Report for the USD-7 Electronic (S) Filtron Company, Inc., July - Aug. '60

208. Reconnaissance System, Eighteenth Monthly Report for the USD-7 Electronic (S) Filtron Company, Inc., Nov. '60

209. Reconnaissance System, Nineteenth Monthly Report for the USD-7 Electronics (S) Filtron Company, Inc., Nov. '60

210. Reconnaissance System, Twenty-First Monthly Report for the USD-7 Electronic (S) Filtron Company, Inc., Feb. '61

211. Receivers and Transmitters, Good Design Practices for Reducing Interference Producing Properties of Vol. II (U) HRB-Singer, Inc., Dec. '59

212. Receivers and Transmitters, Good Design Practices for Reducing Interference Producing Properties of (U) R. B. Marcus, V. M. Ritter, J. G. Taylor, HRB Singer, Inc.

213. Receivers, Bands 1 - 6, RFI Test Report on USD-7 (S) Filtron Company, Inc., April '62

214. Receivers, Bands 1 - 4, Radiated Interference Measurements on USD-7 Scanning (S) Filtron Company, Inc., June '62

215. Receiver, Band 12, RFI Test Report on Scan (S) Filtron Company, Inc., June '62

BIBLIOGRAPHY

216. Receiver Frequency Tuner LRU-2-6, Radio Interference Tests on the Scanner (S) Filtron Company, Inc., Dec. '61

217. Receiver Group, An Interference Study of the Hallicrafters Co. Phase 3 (S) Sprague Electric Company, June '58

2 218. Receiver Performance, Determining True (U) R. B. Marcus, J. G. Taylor, Electronic Industries, Feb. '62

219. Receiver Selectivity, Dynamic (U) Electronics, Feb. '55

220. Receivers, Controlling RFI Susceptibility in (U) Electronic Industries, Sept. '60

221. Relay, Report of Test on Suppression of Interference from DC (U) Canadian Dept. of Transport, Telecommunications Division, Feb. '38

222. Relays, The Dynamics (U) Prof. Charles E. Cameron, Electronic Industries, Sept. '59

223. Report, Trip or Meeting - Location: Hallicrafters (C) Filtron Company, Inc., Feb. '61

224. Report, Trip or Meeting - Location: Hallicrafters (C) Filtron Company, Inc., March '61

225. Report, Monmouth II Monthly Progress (C) Filtron Company, Inc., Feb. '56

226. Report on Progress, Fifteenth Monthly Technical (S) Airborne Instruments Lab., Feb. '60

227. Report, Eighteenth Monthly Technical Progress (S) Airborne Instruments, May '60

228. Report USD-7 for May 1960, Technical Progress (S) Airborne Instruments Labs., June '60

229. Report for July 1960, Technical Progress (S) Airborne Instruments, July '60

230. Research Data Collection and Analysis, Interference Applied - Volume I, Technical Proposal (S) Filtron Company, Inc., Feb. '60

231. Roll Gimble Motor, Preliminary RFI Evaluation of (U) Filtron Company, Inc., April '59

232. Screened Enclosures, Attenuation Measurements Double Shielding - Wire Mesh Cell Type (U) International Electronics Engineering Inc.

233. Seams and Gaskets Used in Shielding Enclosures, and Development of Correctives for Improvement, Evaluation of (U) ASTIA, Document AD-9494

234. Shielded Cables to Reduce Interference, Applying (U) Arnold L. Albin, Filtron Company, Inc.

235. Shielded Rooms, Practical Aspects in Evaluating (U) James C. Klouda, Electro-Technology, June '61

236. Shielding Considerations for BMEWS, General (U) Filtron Company, Inc., Oct. '58

237. Shielding Considerations, Fundamental - Lecture Notes (U) Filtron Company, Inc., Feb. '59

238. Shielding for the Suppression of Radio Interference (U) Canadian Dept. of Transport, Telecommunications Div., Feb. '53

239. Shielding with Knitted Metal Mesh Gaskets, RFI (U) Military Systems Design, May-June '62

240. Side Lobe Reduction Techniques, Study and Development of (S) ASTIA, Jan. '59

241. Solid-State Digital Computer Model, Radio Interference and Susceptibility Study of a (U) J. A. Harder, R. L. Powers IBM Corp.

242. Solid-State Rectifier Devices as Radio Noise Suppression Components, Final Report and Investigation of (U) Worcester Polytechnic Institute

243. Spurious Response and the Generation of Spurious Radiation, An Investigation of Fundamental Circuit Factors Which Influence (U) Electro-Mechanics Co., Dec. '55

244. Soils and Concrete for BMEWS Site III, Study of Propagation Characteristics of (C) Filtron Company, Inc., May '60

245. Single Interference Frequency, Suppressing (U) Electronic Industries, May '61

246. Suppression Components and Networks, Effective Utilization - Engineering Memorandum (U) RCA Engineering Products Div., Feb. '56

247. Suppression Techniques, Radio Interference (U) Coles Signal Lab, Signal Corps Engineer Labs, Nov. '53

248. Switching Centers, Shielding Specification Requirements, Preferred Construction Practice, and Methods of Test for North Electric Company Electronic (S) Filtron Company, Inc., Oct. '60

249. Systems, Electrical Noise Reduction in Electronic (U) Robert L. Wendt, Sperry Gyroscope Company

250. Systems Design, Consider Interference in (U) Rocco F. Ficcki, Electronic Industries, Mar. '60

251. Systems Interference Evaluation Project (Call-7) (S) Filtron Company, Inc., July '60

252. System RFI Program, First Monthly Report for Call-7 (S) Filtron Company, Inc., Oct. '60

253. Technical Summary Report - Contract No. NAS 8-846 (U) Filtron Company, Inc.

254. Techniques, Accuracy for Improvement (C) Federal Scientific Corp., Feb. - April '59

255. Telemetering Package, Radio Interference Evaluation of Bendix FM/FM (C) Filtron Company, Inc., Jan. '57

256. Test Equipment, Interference Evaluation of Atlas Ground (C) Filtron Company, Inc., June '57

257. Test and Measurements Program, Down Range (S) Filtron Company, Inc., May '60

258. Test Plan (RFI), Review of Bendix-Friez BX 7 (C) Filtron Company, Inc., May '61

259. Test Program, Alaskan (C) Filtron Company, Inc., Oct. '60

260. Testing ORC 68 (T) MOD I, Environmental and Interference - Appendix I (U) Hallicrafters, June '59

BIBLIOGRAPHY

261. Thermistors, Voltage Suppression with PTC (U) A. Hemel, Electronic Design, Aug. '62

262. Toroidal Inductor Design, Simplified Approach to (U) Tele-Tech, Jan. '52

263. Transmission Lines, Report of a Review of Past and Current Work on Electromagnetic Interference Voltage from Electric Power (U) Filtron Company, Inc., Nov. '55

264. Transmitter Spurious Radiation, New Components for Reduction of (U) Electro Mechanics Co., June to Sept. '61

265. Transmitter Spurious Radiation, New Components for Reduction of (U) Electro-Mechanics Company, Sept. to Dec. '62

266. Transmitters RFI-Free, Making (U) C. E. Blankely, R. N. Bailey, Electronic Industries, March '60

267. UHF Equipments, Final Development Report on Mutual Environmental Tests of, Phase IV (C) Collins Radio Company, Mar. '56

268. USD-7 Band 10, RFI Test Report on (S) Filtron Company, Inc., June '62

269. USD-7 Band 11, RFI Test Report on (S) Filtron Company, Inc., June '62

270. USD-7 Band 14 LRU 2-14 and LRU 2-96 (AIL-E) RFI Test Report on (S) Filtron Company, Inc., May '62

271. USD-7 Wide Open Receiver (AIL-E), Radio Interference Tests on, TM-1065-18 (S) Filtron Company, Inc., Oct. '61

272. Vulnerability, Division Radio Central, Project Monmouth III (C) U.S. Army Signal R & D Lab, June '58

273. Warning System, Interference Analysis in the Ballistic Missile Early - Task 7.0 - Final Report (S) Filtron Company, Inc., July '59

INDEX

INDEX

A

INDEX

B

INDEX

C

INDEX

D

INDEX

E

INDEX

F

F

G

P

Q

R

INDEX

T

U

V

W

Z

The editor wishes to gratefully acknowledge his indebtedness to the many authors, scientists, engineers and publishers who kindly granted permission to reproduce their graphs, charts and material.

The editor has been dependent upon many sources of information, reports, documents, technical papers, and publications for the material contained in these two volumes. Without the permission of those listed below, and others, it could not have been possible. If omissions have occurred, it has been unintentional.

LIST OF CONTRIBUTORS

Publication	Publisher
Applications and Industry	Institute of Electrical and Electronics Engineers
Bell Laboratories Record	Bell Telephone Laboratories
Communication and Electronics	Institute of Electrical and Electronics Engineers
CQ	Cowan Publishing Corp.
Electrical Engineering	Institute of Electrical and Electronics Engineers
Electronics and Communications	Southam Business Publications
Electronic and Radio Engineer	Iliffe and Sons, Ltd., London, England
Electronics	McGraw - Hill, Inc.
Electronic Design	Hayden Publishing Company, Inc.
Electronic Engineering	Morgan Bros., Ltd., London, England
Electronics Engineering of Canada	Maclean - Hunter, Ltd., Toronto, Canada
Electronic Industries and Tele-Tech	The Chiltron Co.
Electronic Products	United Technical Publications
Electronic Progress	Raytheon Manufacturing Co.
Electronic Technology	Iliffe & Sons, Ltd., London, England
Electronic Week	Hayden Publishing Company, Inc.
Electro - Technology	C-M Technical Publications Corp.
General Electric SCR Manual	General Electric Company
General Electric Review	General Electric Company
General Motors Engineering Journal	General Motors Corporation
General Radio Experimenter	General Radio Co.
Hewlett-Packard Journal	Hewlett-Packard Co.
IEEE Proceedings	Institute of Electrical and Electronics Engineers
Illuminating Engineer	Illuminating Engineering Society
Interference Reduction Manuals	U. S. Air Force, U. S. Army, U. S. Navy
IRE Proceedings	Institute of Electrical and Electronics Engineers
Microwaves	Hayden Publishing Company, Inc.
Missiles and Rockets	American Aviation Publications, Inc.
News	International Rectifier Co.
The Notebook	Boonton Radio Corp.
Philips Technical Review	N.V. Philips Gloeilampen-fabrieken, Netherlands
Proceedings of the Yearly Conference on Radio Interference Reduction and Electromagnetic Compatibility	Armour Research Foundation
RFI / EMC	General Electric Company
The Radio and Electronic Engineer	The Institution of Electronic and Radio Engineers, London, England
Radio Engineering	Pergamon Press
Radio Interference Reduction in Aircraft Systems	Filtron Company, Inc.
SAE Journal	Society of Automotive Engineers
Sperry Engineering Review	Sperry Rand Corp.
Sylvania Technologist	Sylvania Electric Products
Technical Publications	Allied Research Products, Inc.
Technical Publications	Perfection Mica Co.
Technical Publications	Technical Wire Products, Inc.
Technical Reports, Manuals and Publications	Filtron Company, Inc.
Telecommunications	Pergamon Press
Westinghouse Engineer	Westinghouse Electric Corp.
Wireless Engineer	Iliffe and Sons, Ltd., London, England